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Masaki et al.

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(54) **MOTOR CONTROL DEVICE**

JP A 11-150983 6/1999

(75) Inventors: **Ryoso Masaki**, Hitachi (JP); **Satoru Kaneko**, Urizura-machi (JP); **Sanshiro Obara**, Tokai-mura (JP); **Mitsuyuki Hombu**, Hitachinaka (JP); **Yoshimi Sakurai**, Hitachioota (JP)

(73) Assignee: **Hitachi, Ltd.**, Tokyo (JP)

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(30) **Foreign Application Priority Data**

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(51) **Int. Cl.⁷** **H02P 6/18**; H02P 21/00

(52) **U.S. Cl.** **318/721**; 318/254; 318/701

(58) **Field of Search** 318/138, 254, 318/439, 700, 701, 720, 721, 722, 724

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Primary Examiner—Bentsu Ro

(74) *Attorney, Agent, or Firm*—Crowell & Moring LLP

(57)

ABSTRACT

In an arrangement having an electric power inverter 2 which applies a voltage to an AC motor 1 and a control unit 4 which calculates a voltage command value applied in PWM signals, detection use voltages V_{up} , V_{vp} and V_{wp} are applied in synchronism with PWM signal generation use carrier waves in the control unit 4 as well as motor currents i_v and i_w are detected in response to current detection pulses P_d in synchronism therewith. In a magnetic pole position detection unit 12, counter electromotive forces of the synchronous motor 1 are estimated from a relationship between the detection use voltages V_{up} , V_{vp} and V_{wp} and current difference vectors of the motor current to determine a magnetic pole position θ through calculation. The determined magnetic pole position θ is inputted to coordinate conversion units 8 and 11 and a motor speed detection unit 13. Thereby, a motor control device having a control characteristic of a good response property with no magnetic pole position sensor is provided for a control system of a synchronous motor.

13 Claims, 17 Drawing Sheets

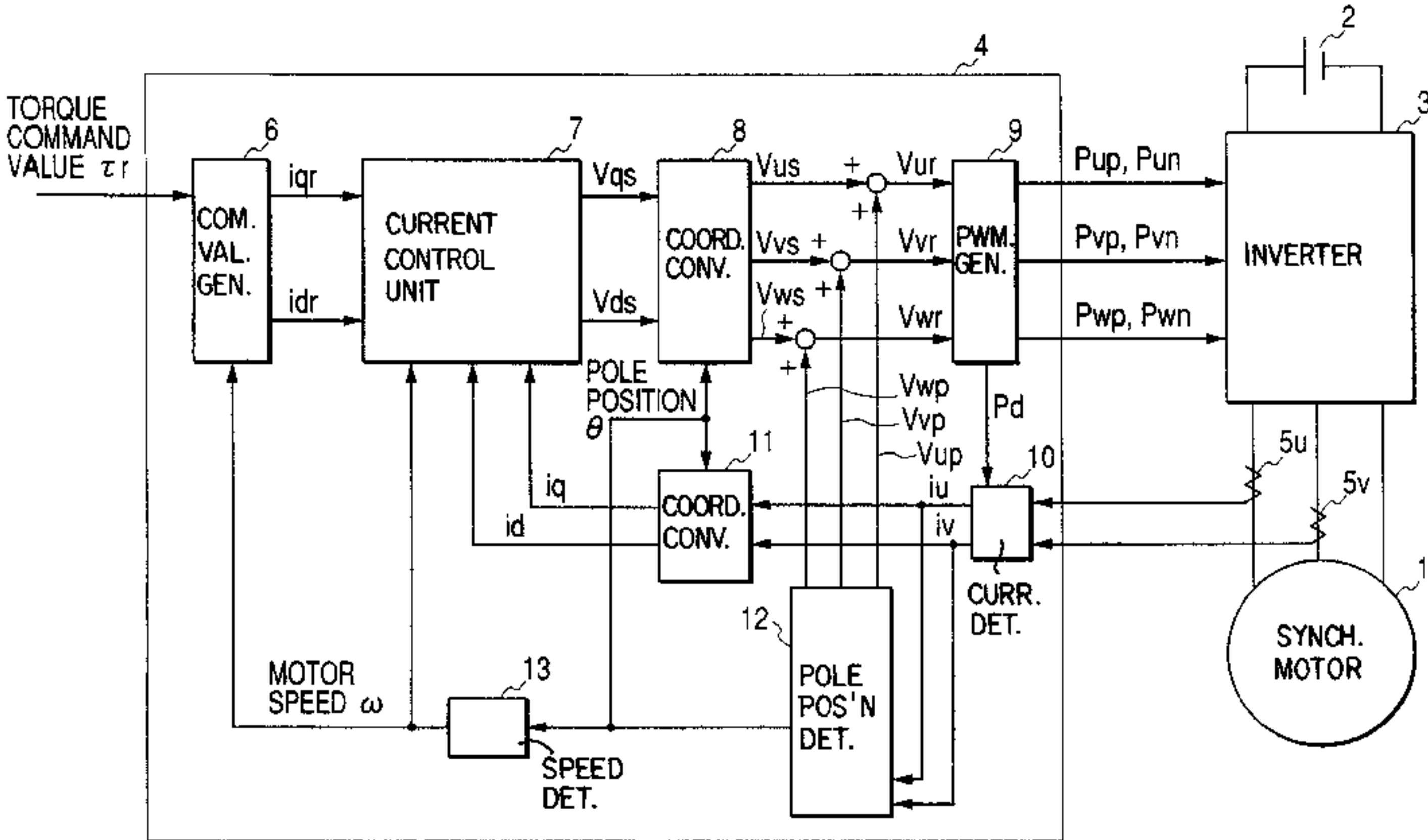


FIG. 1

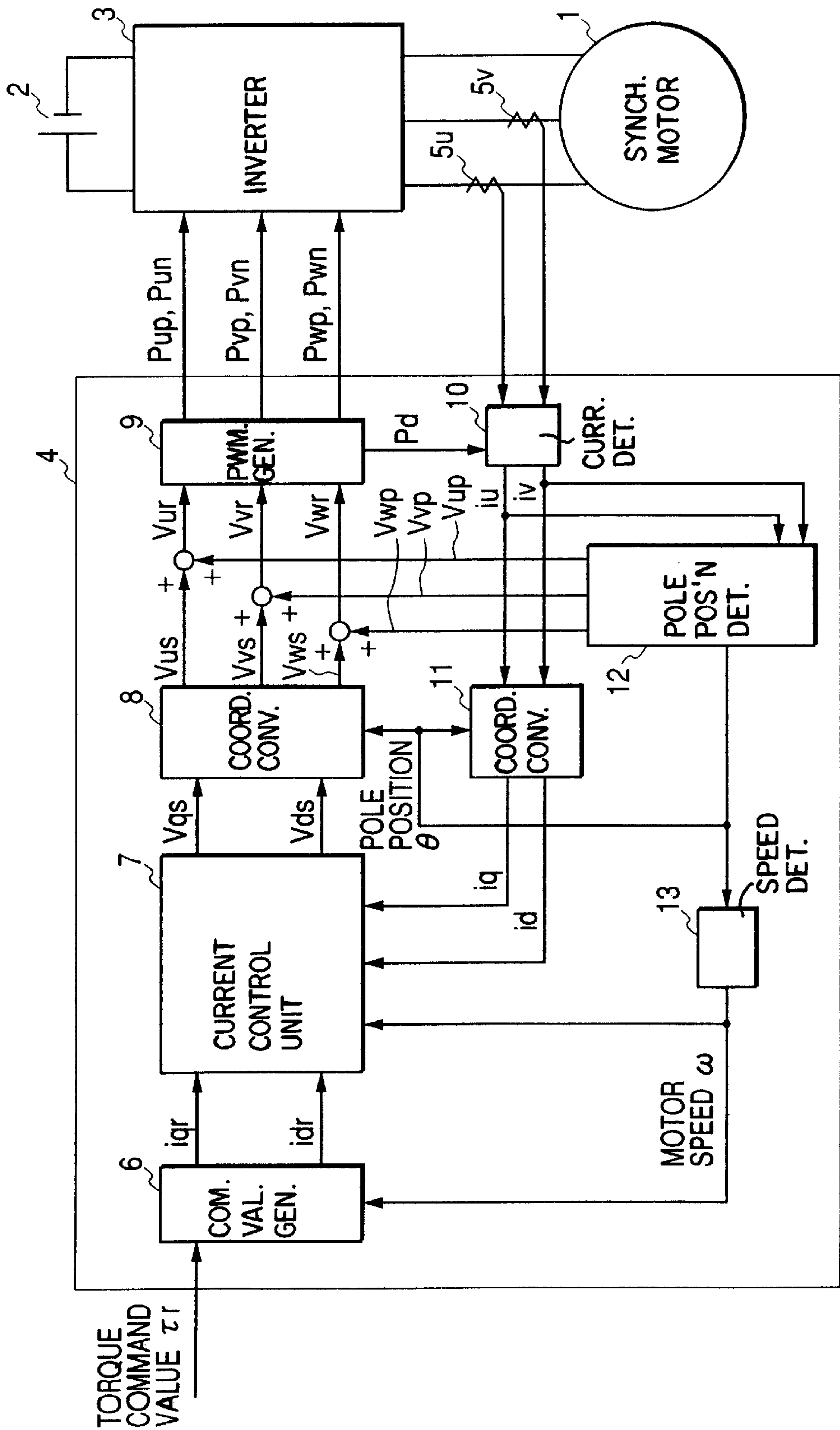


FIG. 2

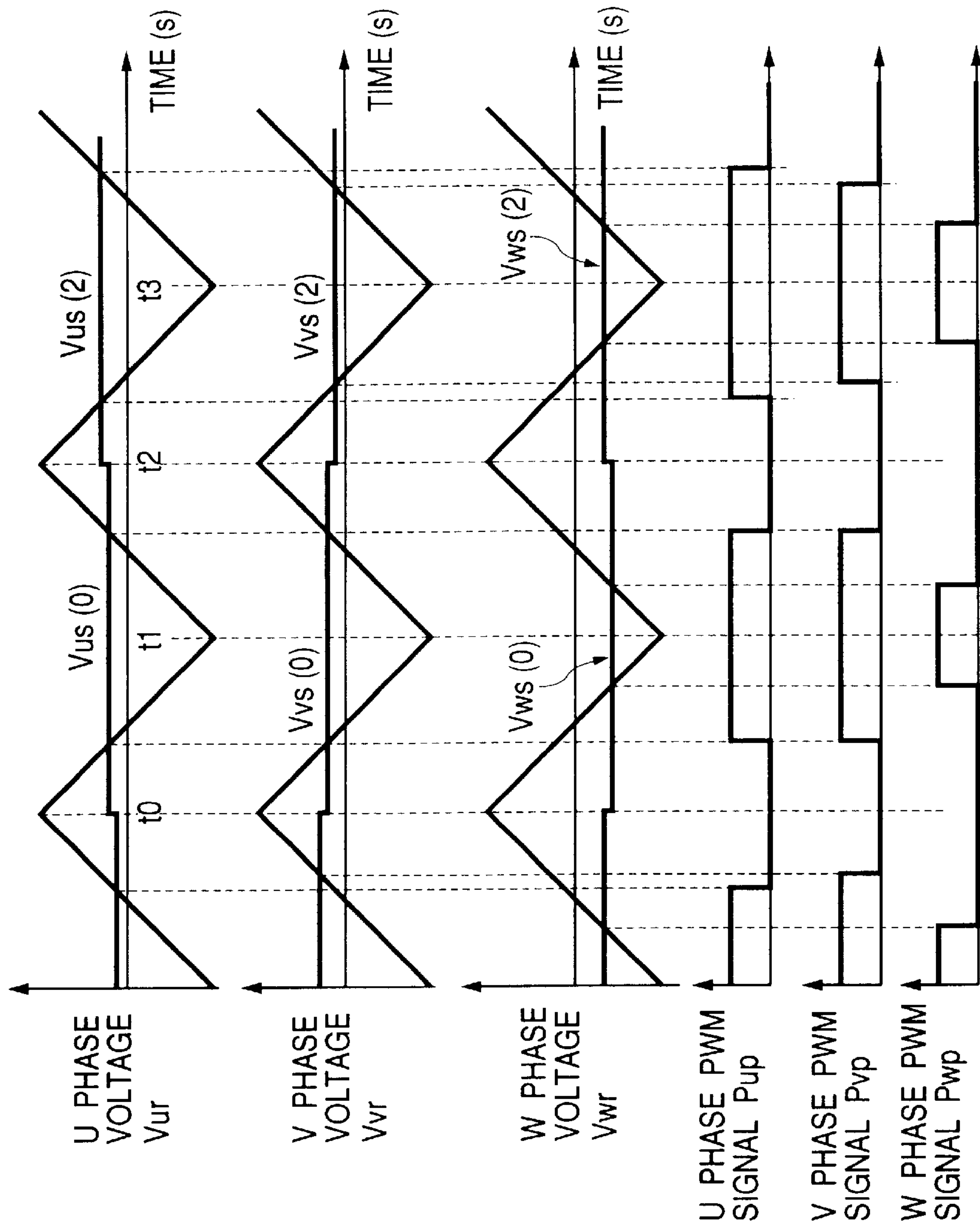


FIG. 3

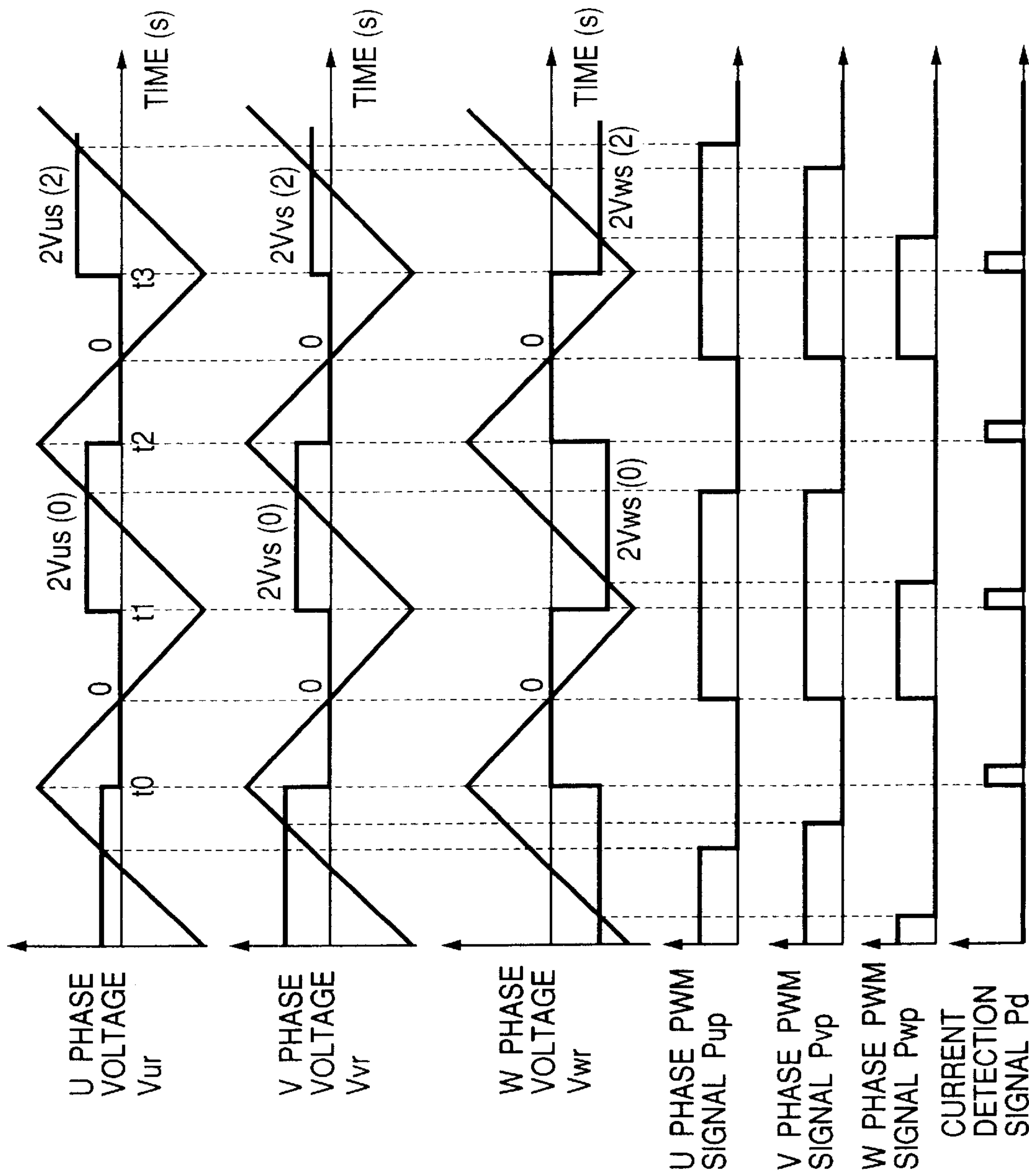


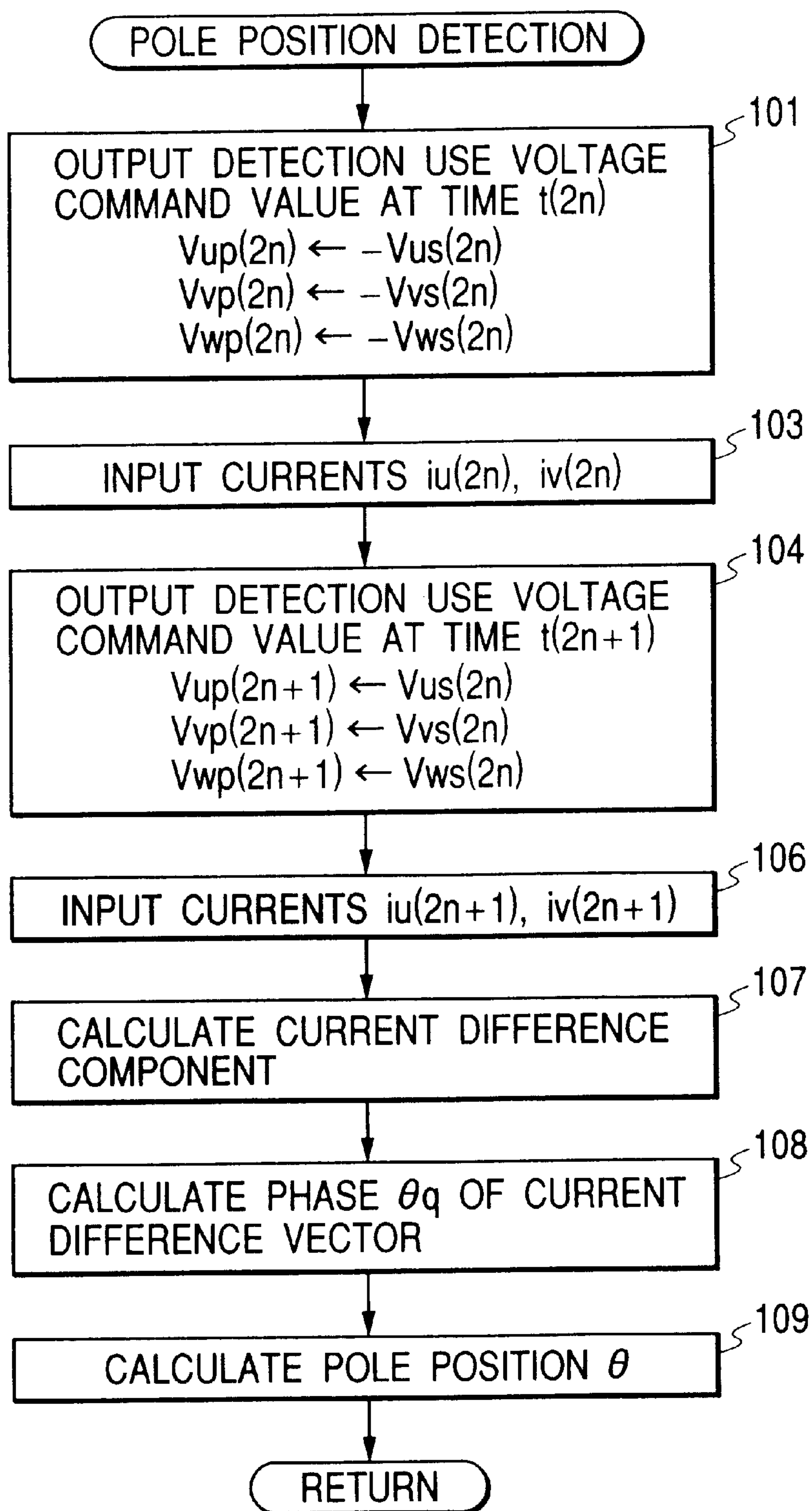
FIG. 4

FIG. 5

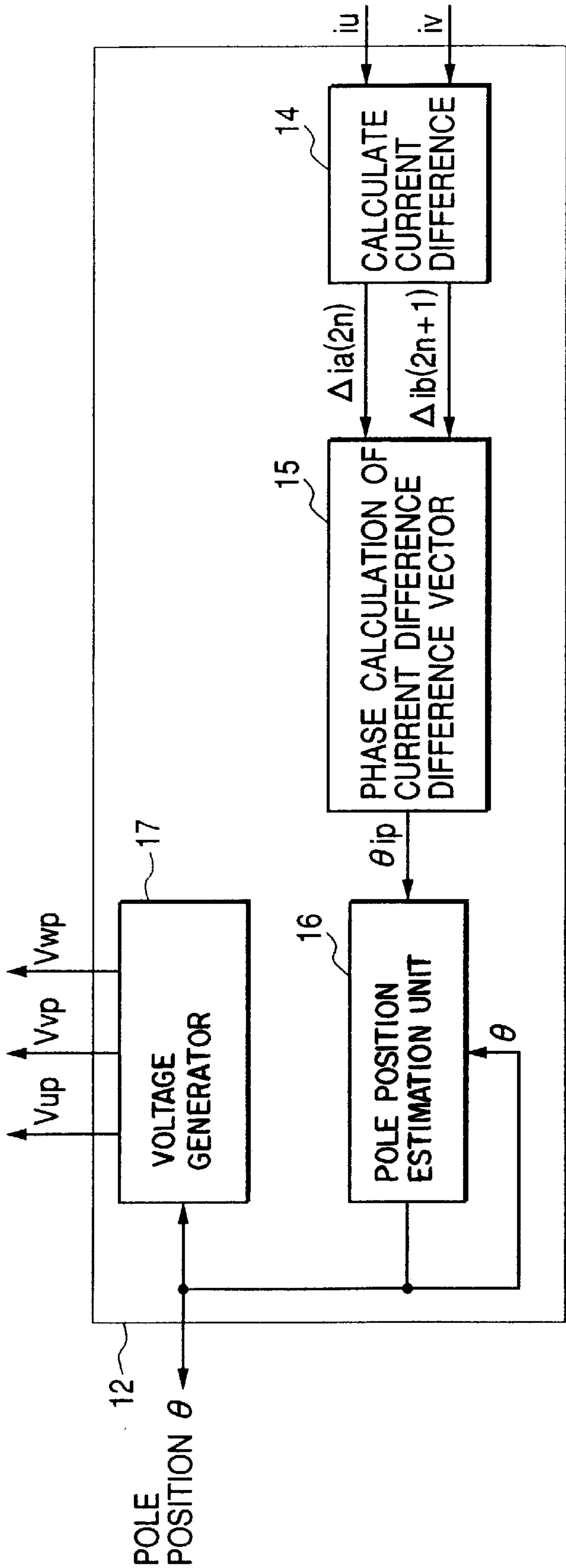


FIG. 6A

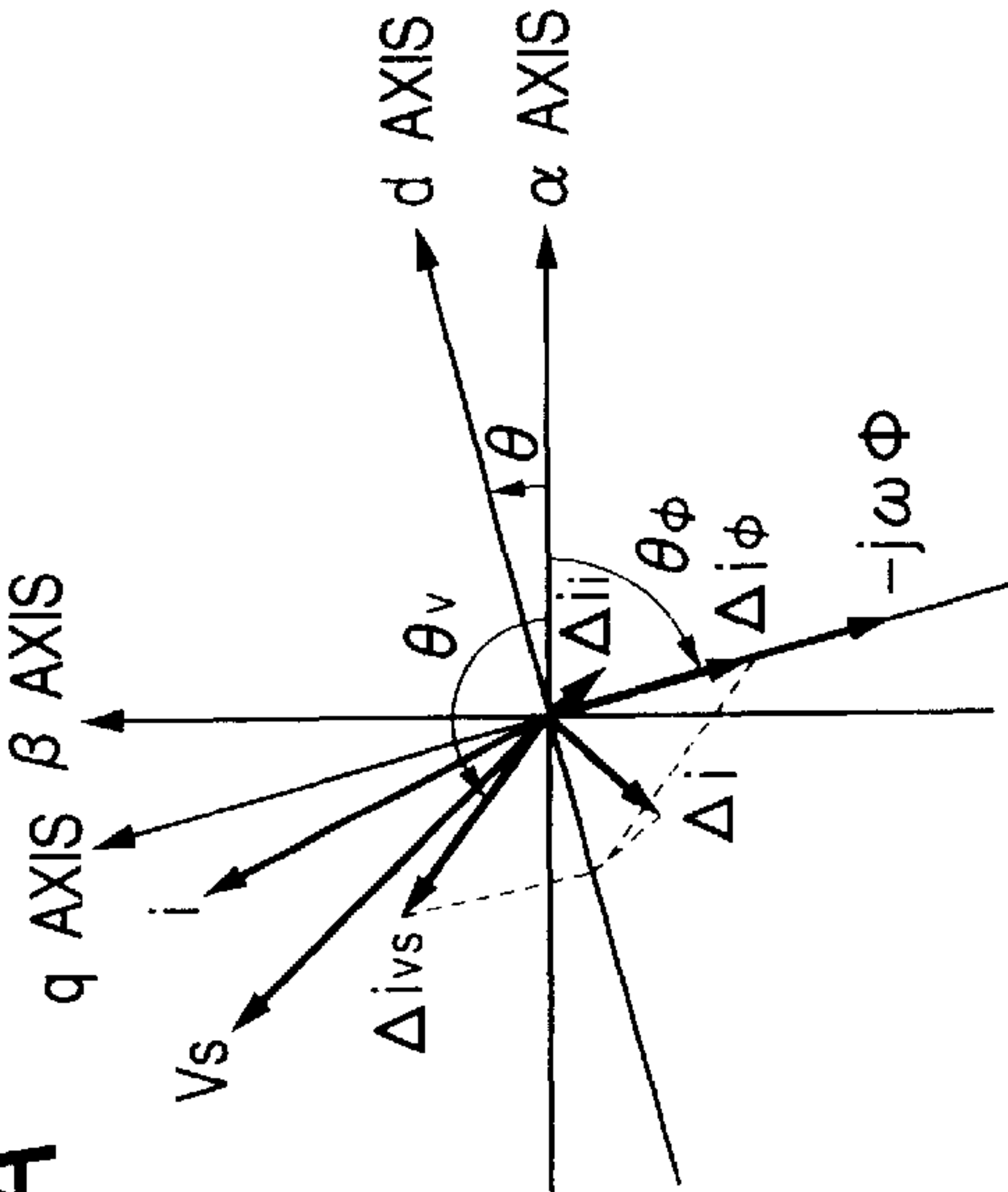


FIG. 6B

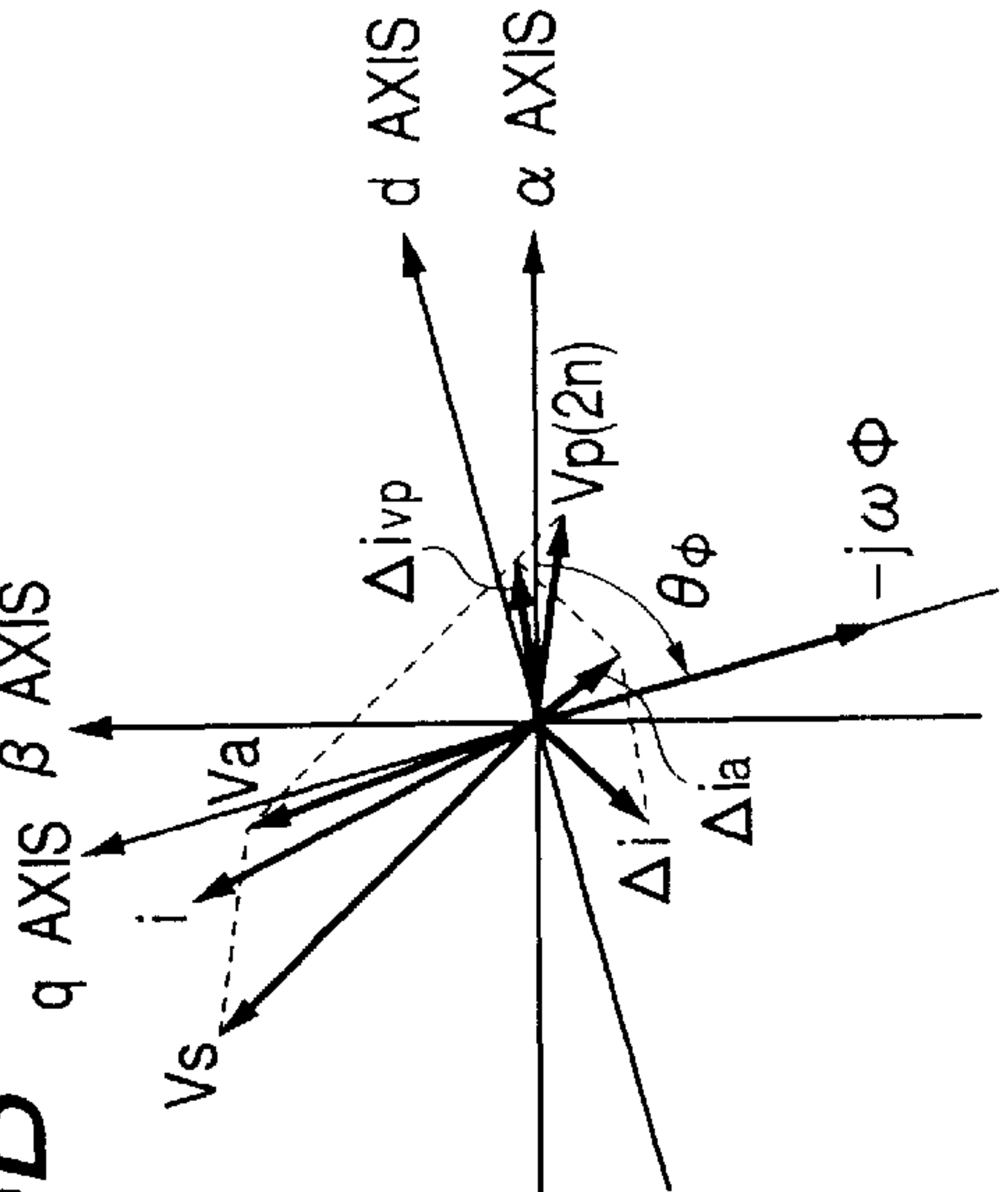


FIG. 6C

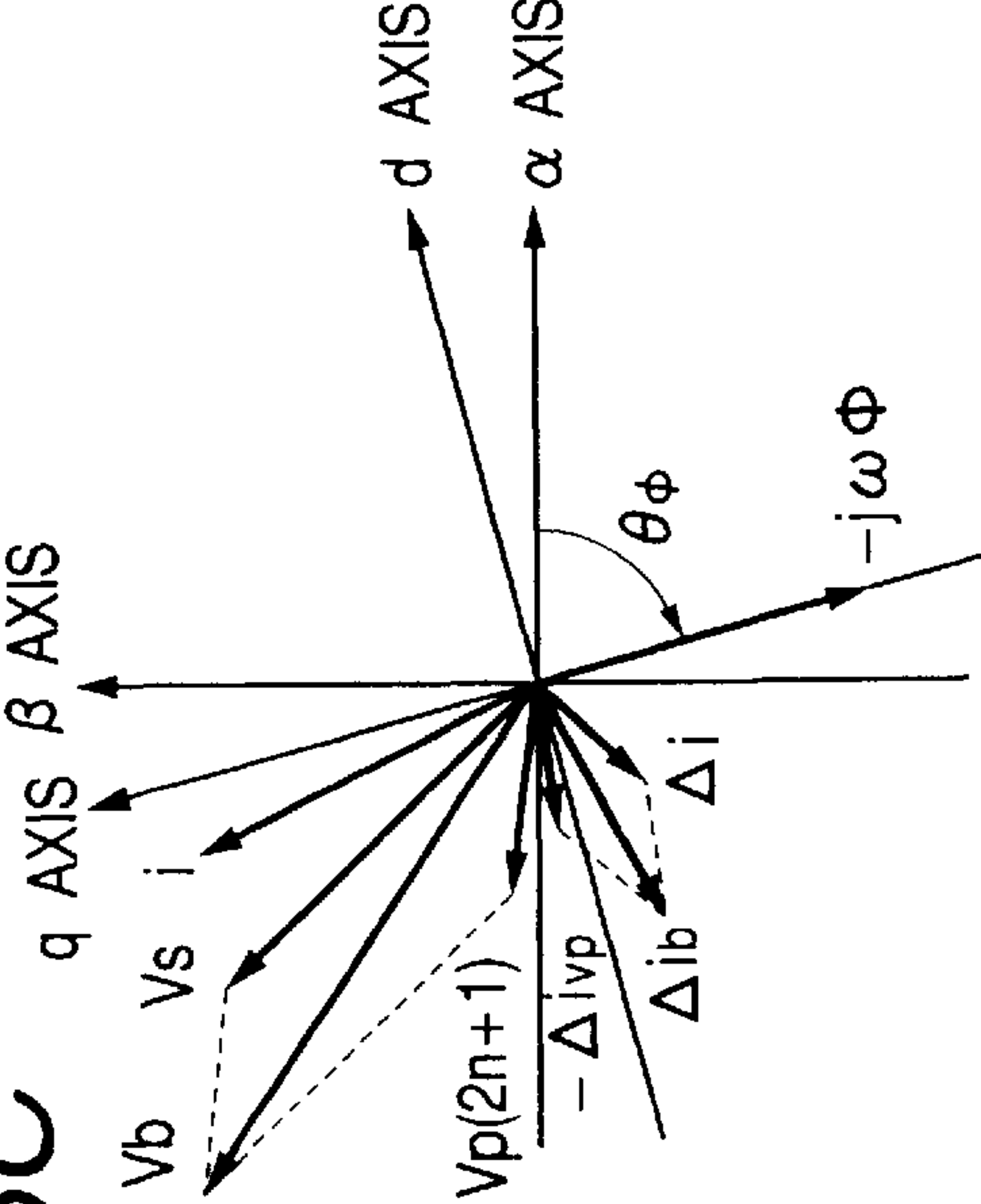


FIG. 6D

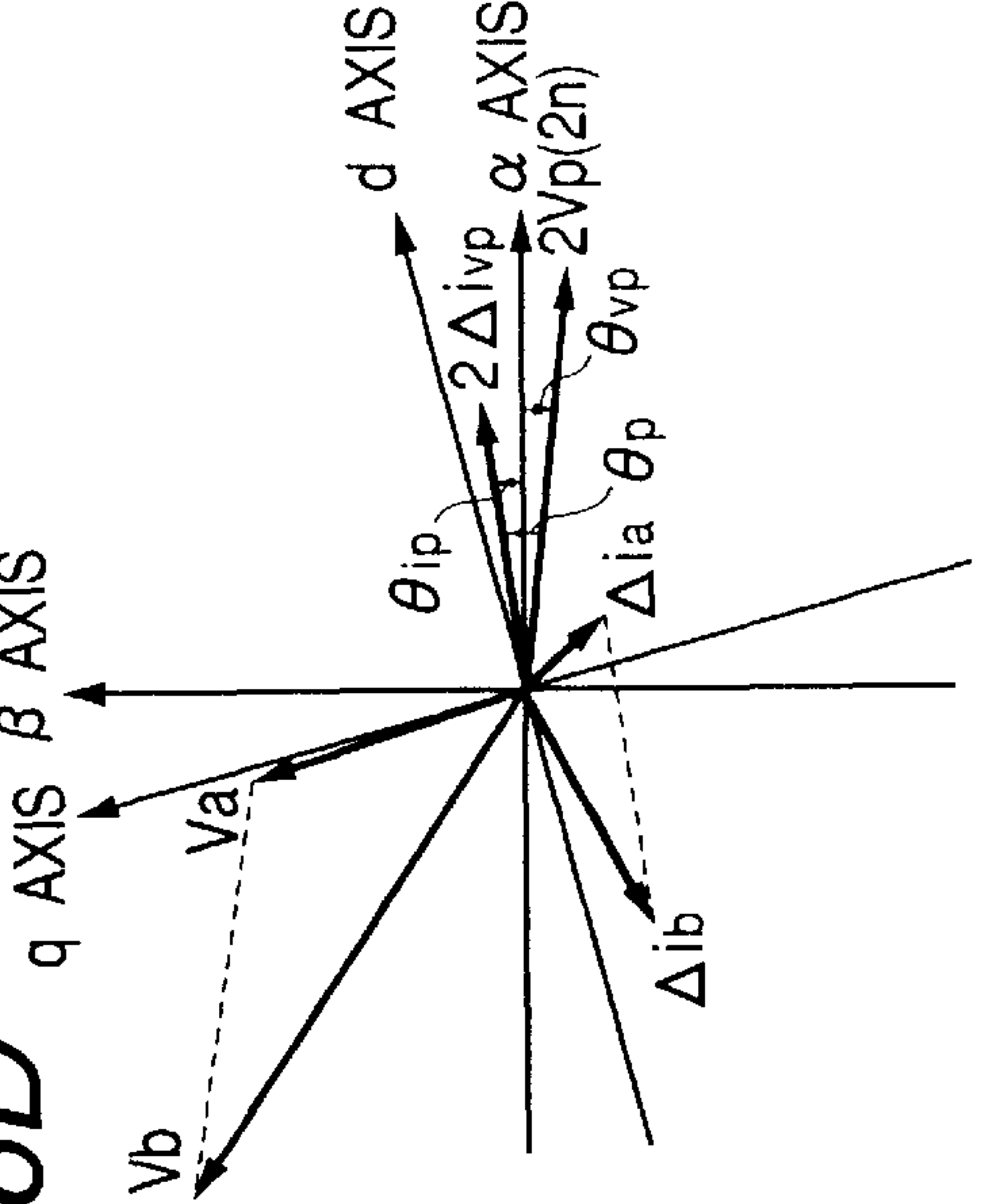


FIG. 7

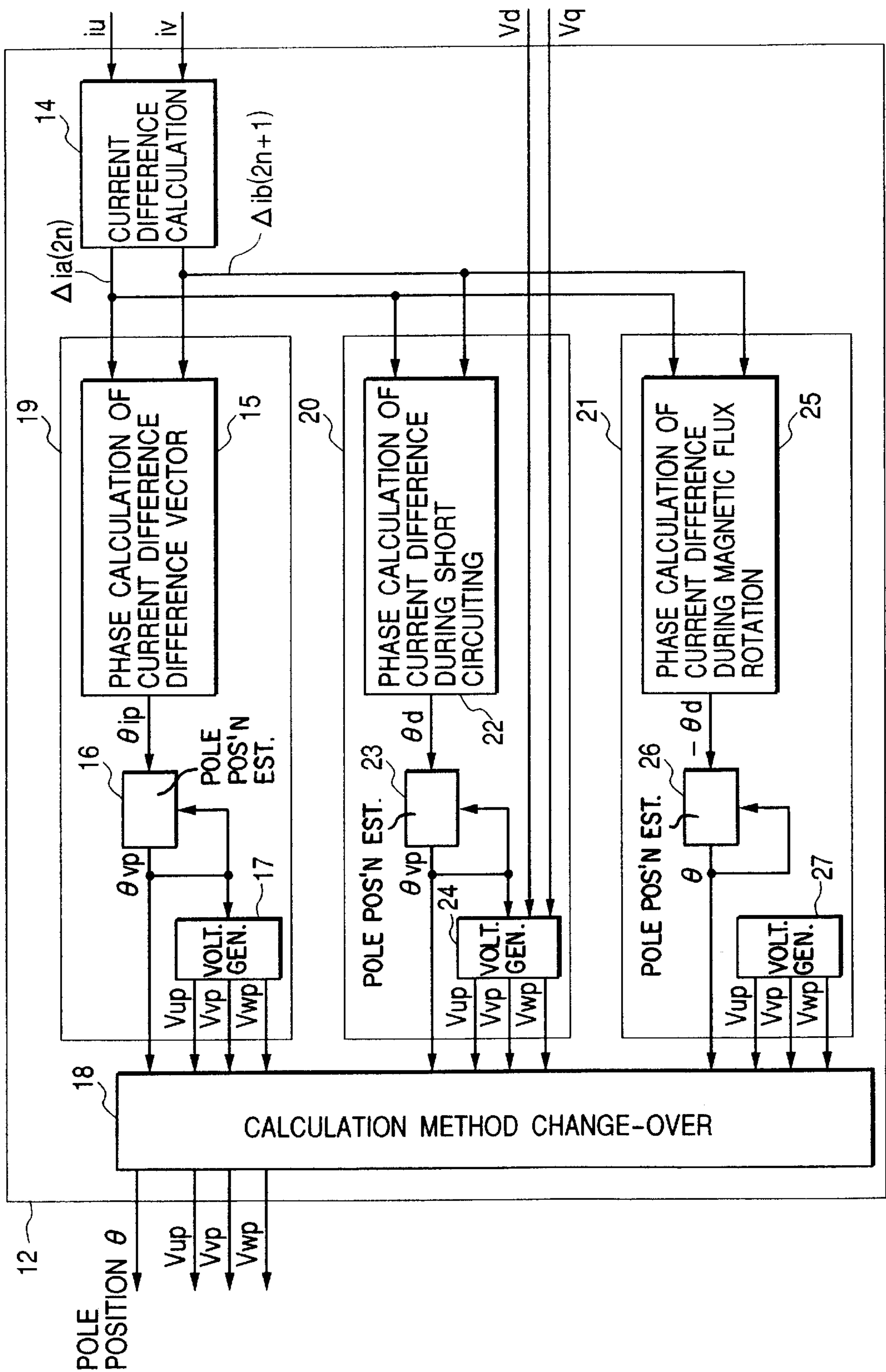


FIG. 8

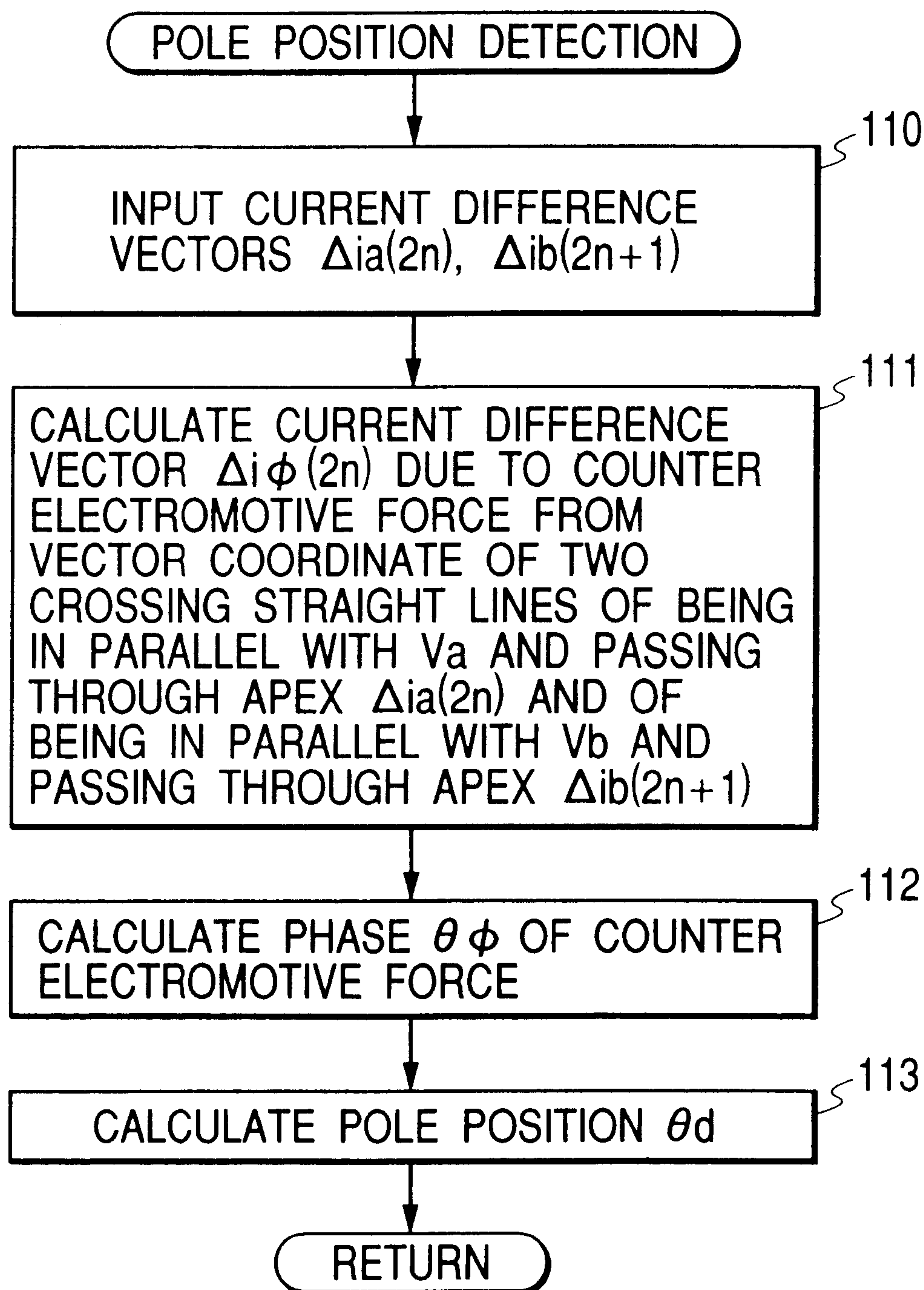


FIG. 9A

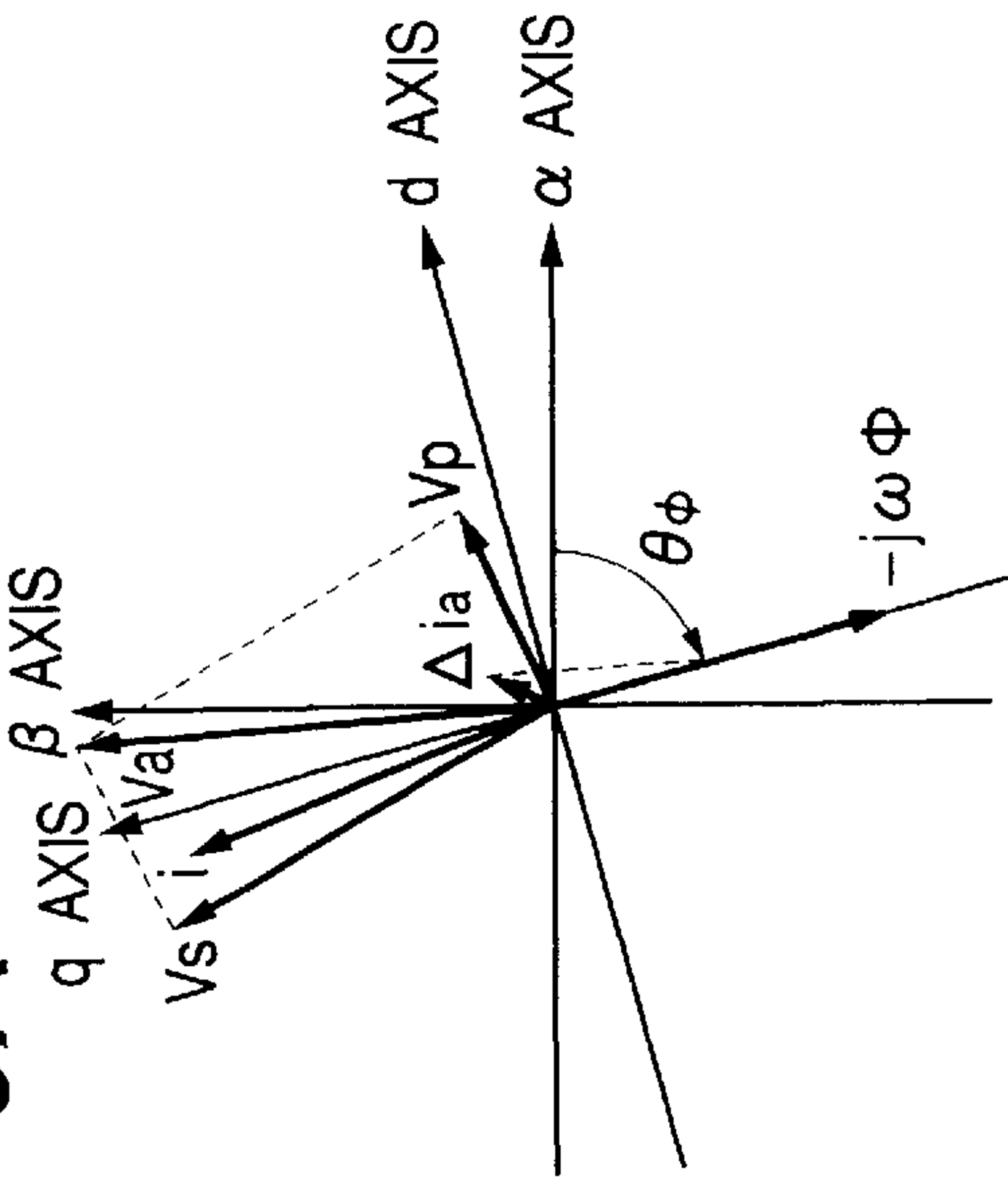


FIG. 9B

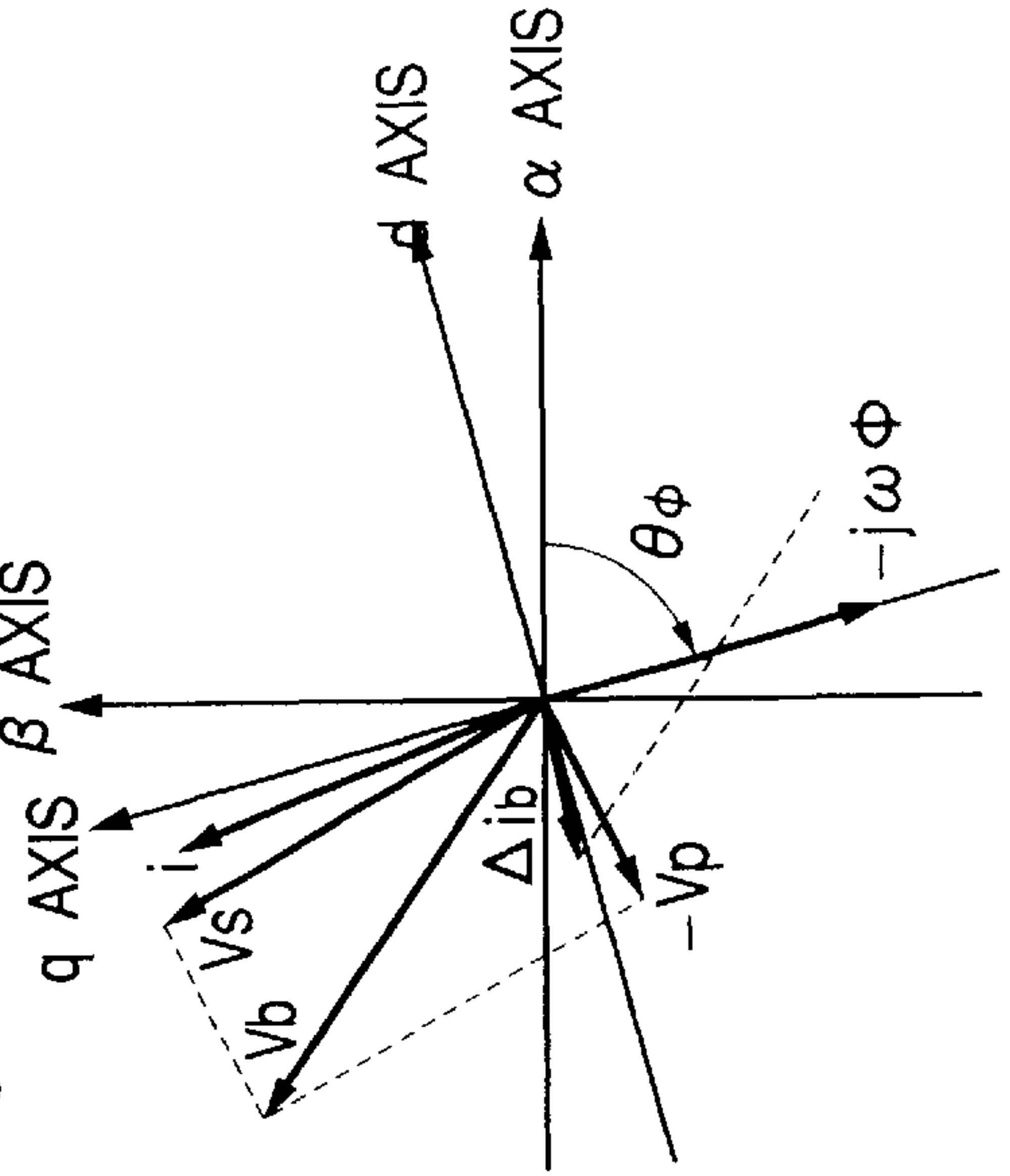


FIG. 9C

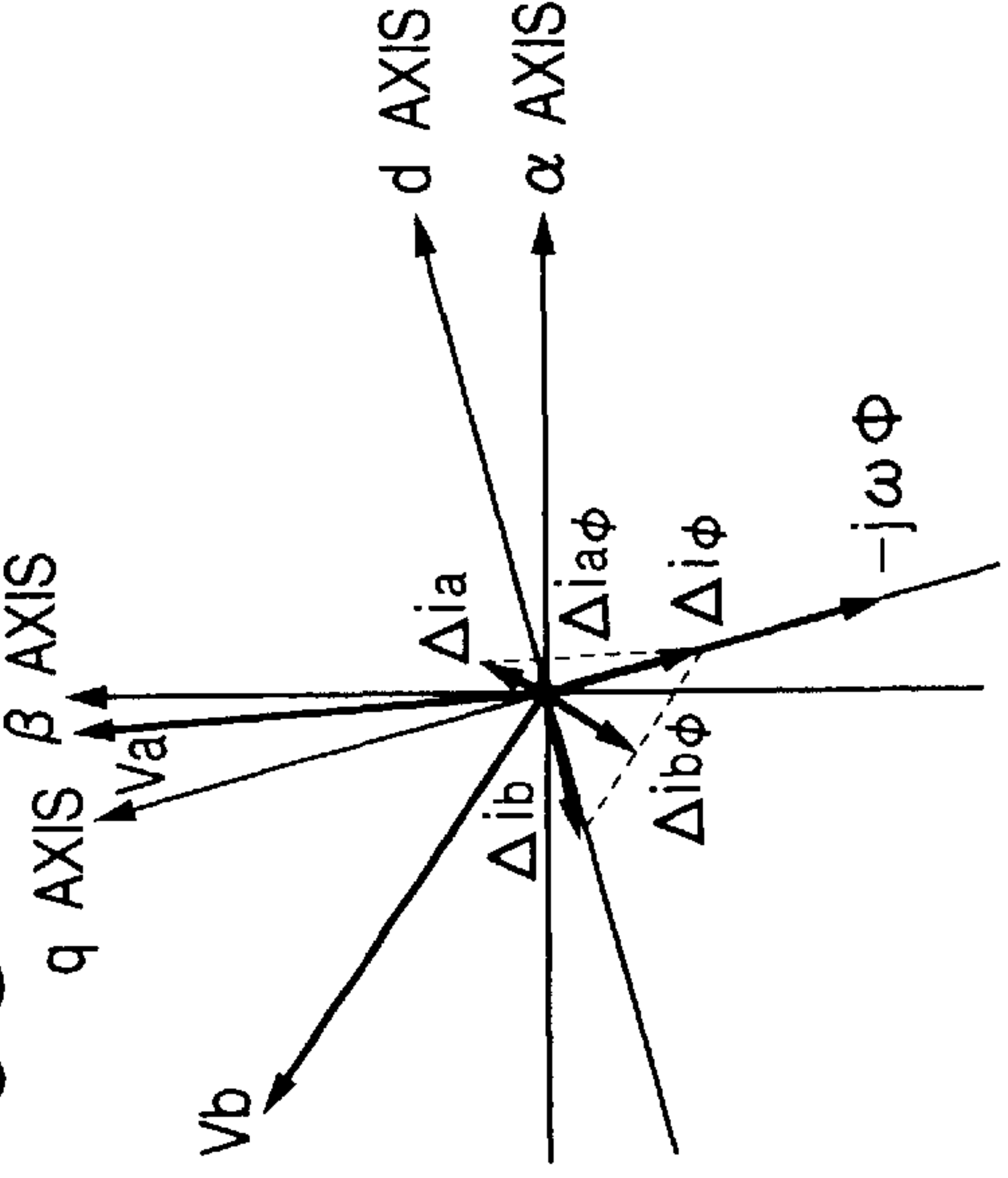


FIG. 10

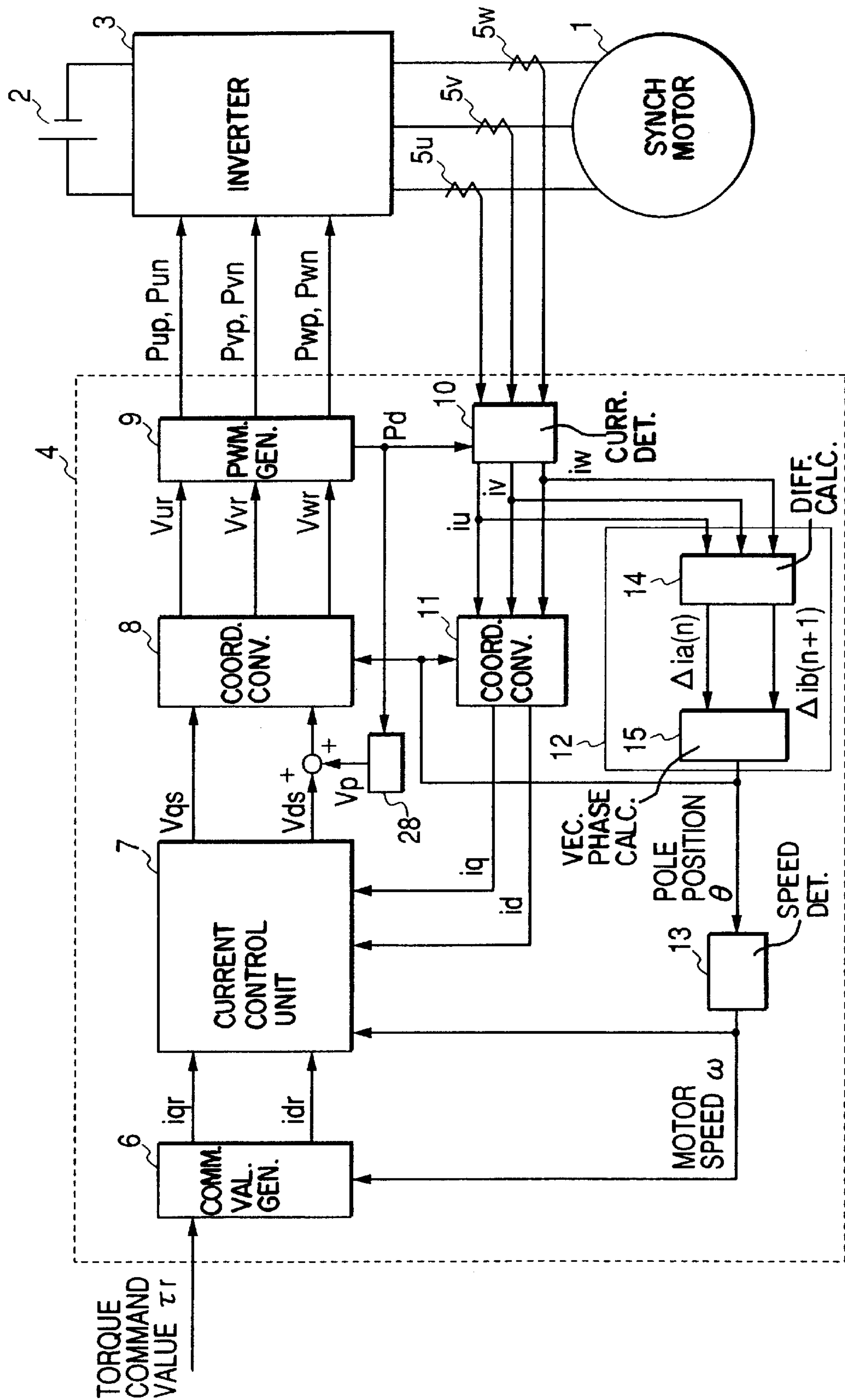


FIG. 11

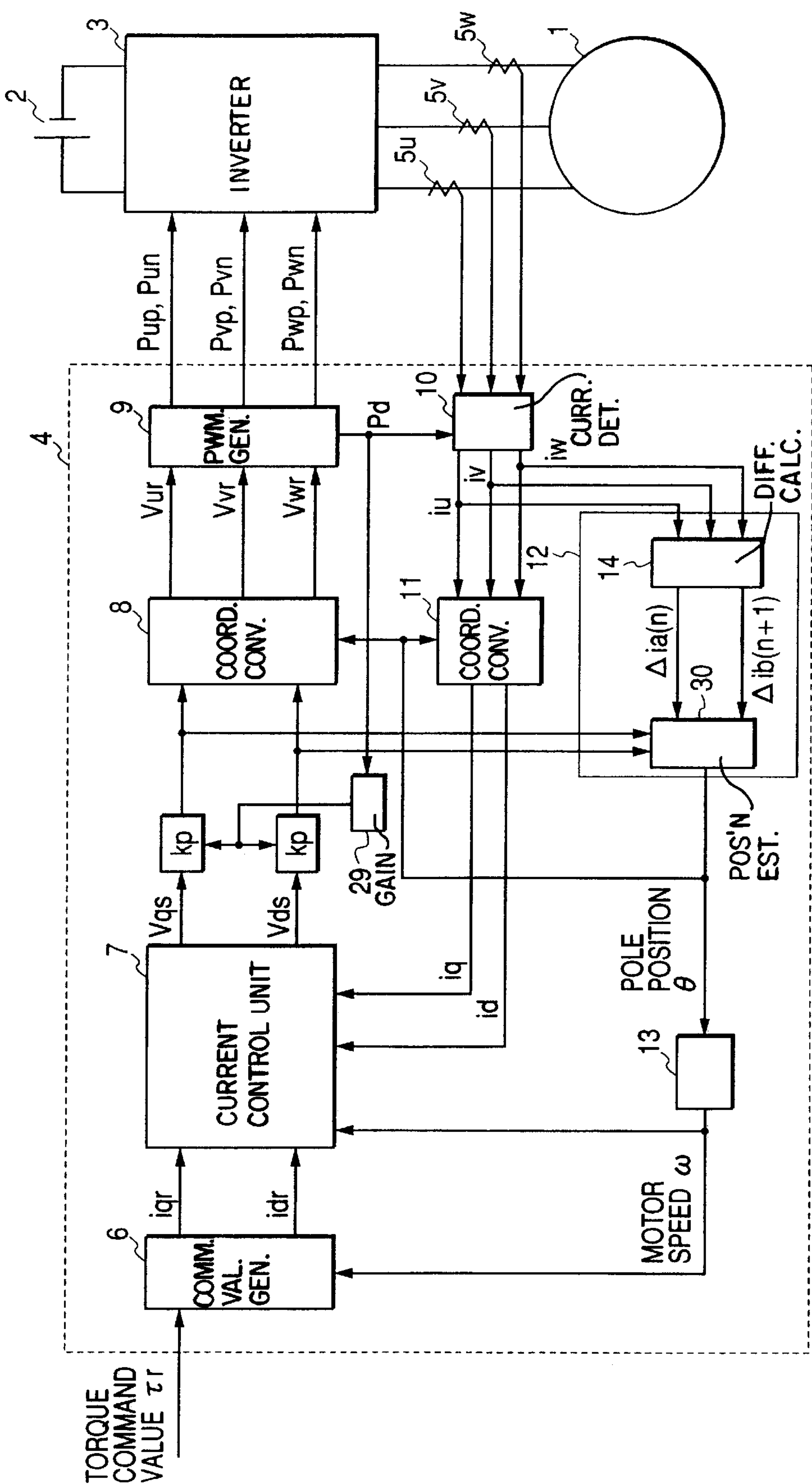


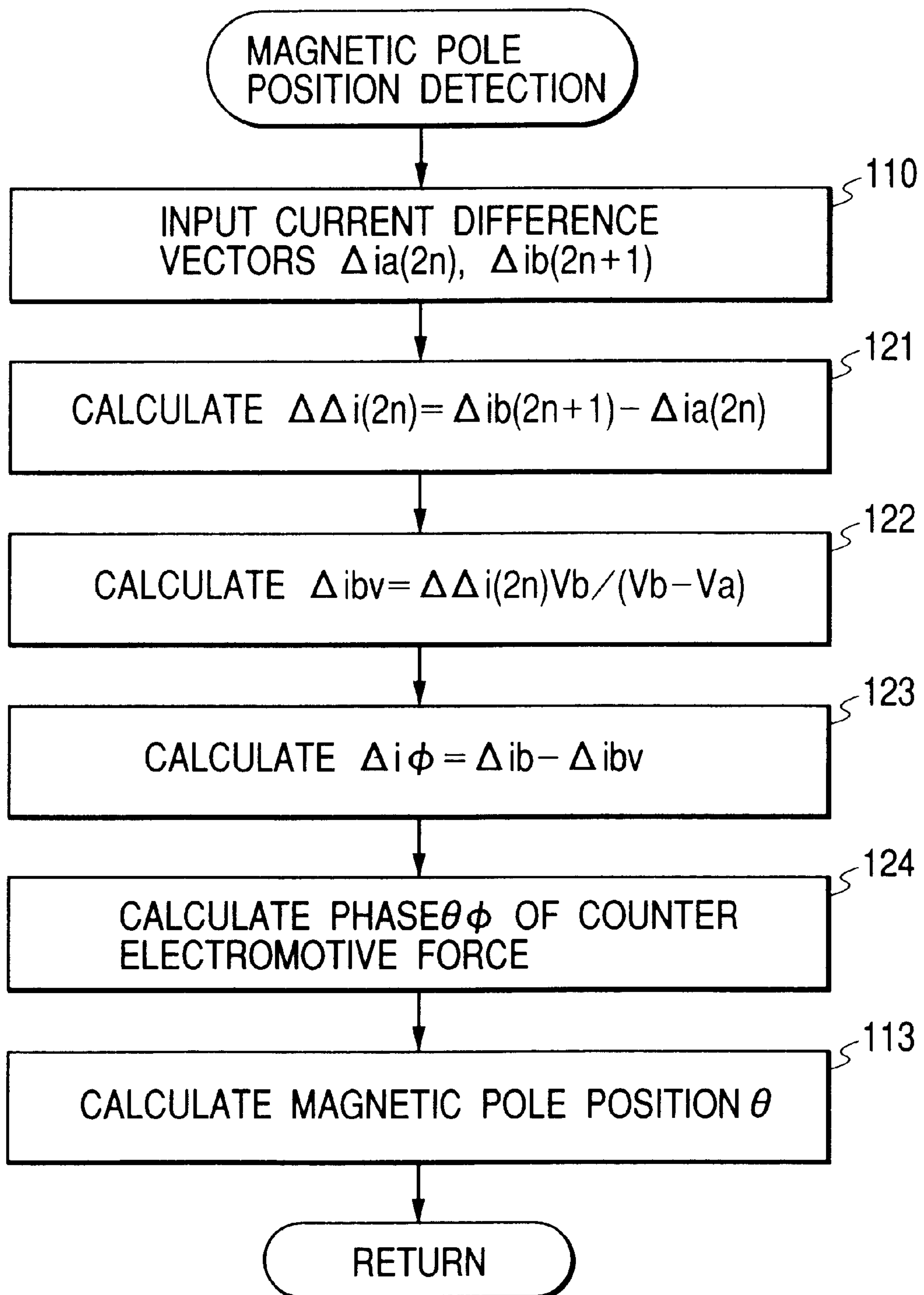
FIG. 12

FIG. 13A

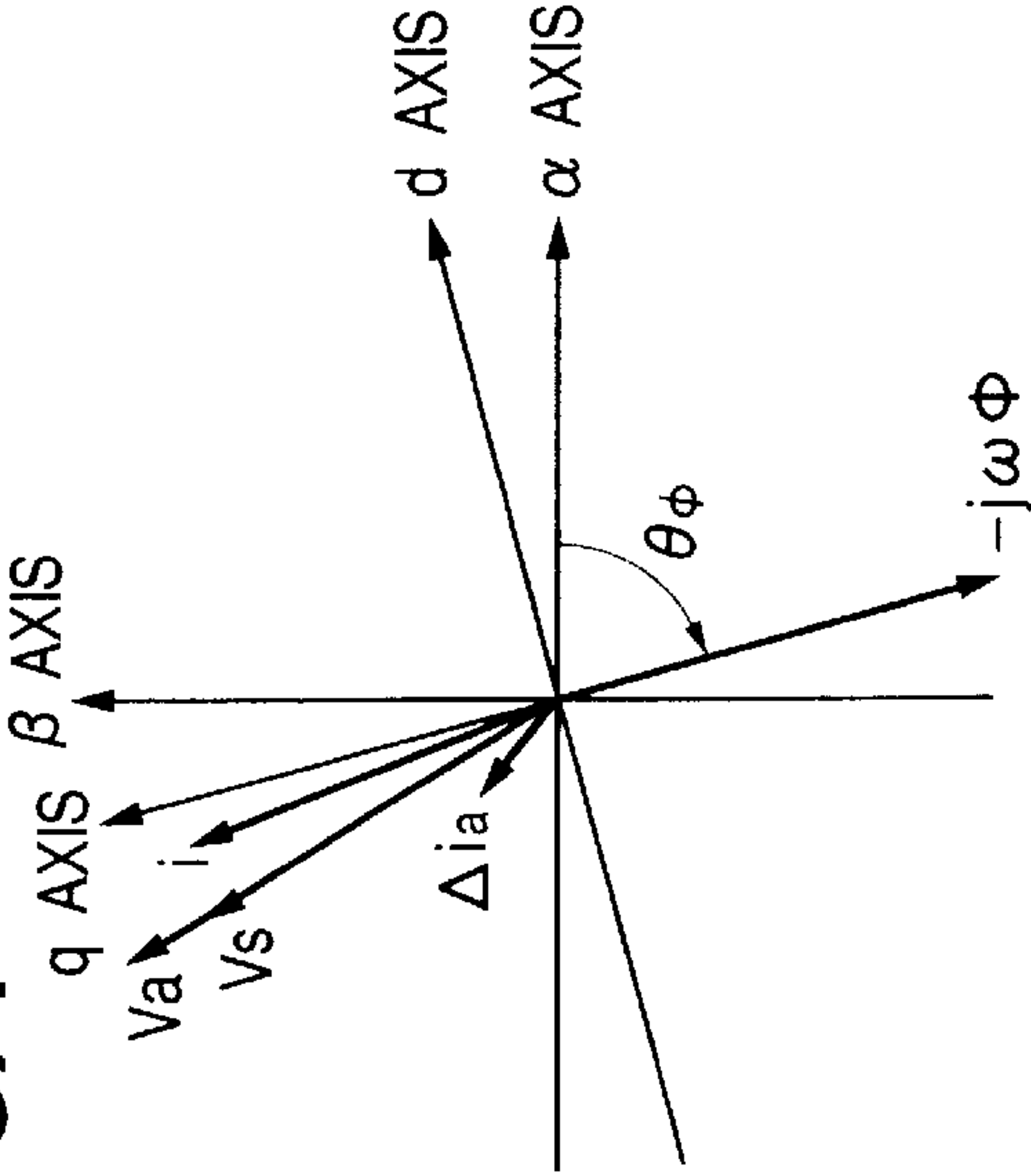


FIG. 13B

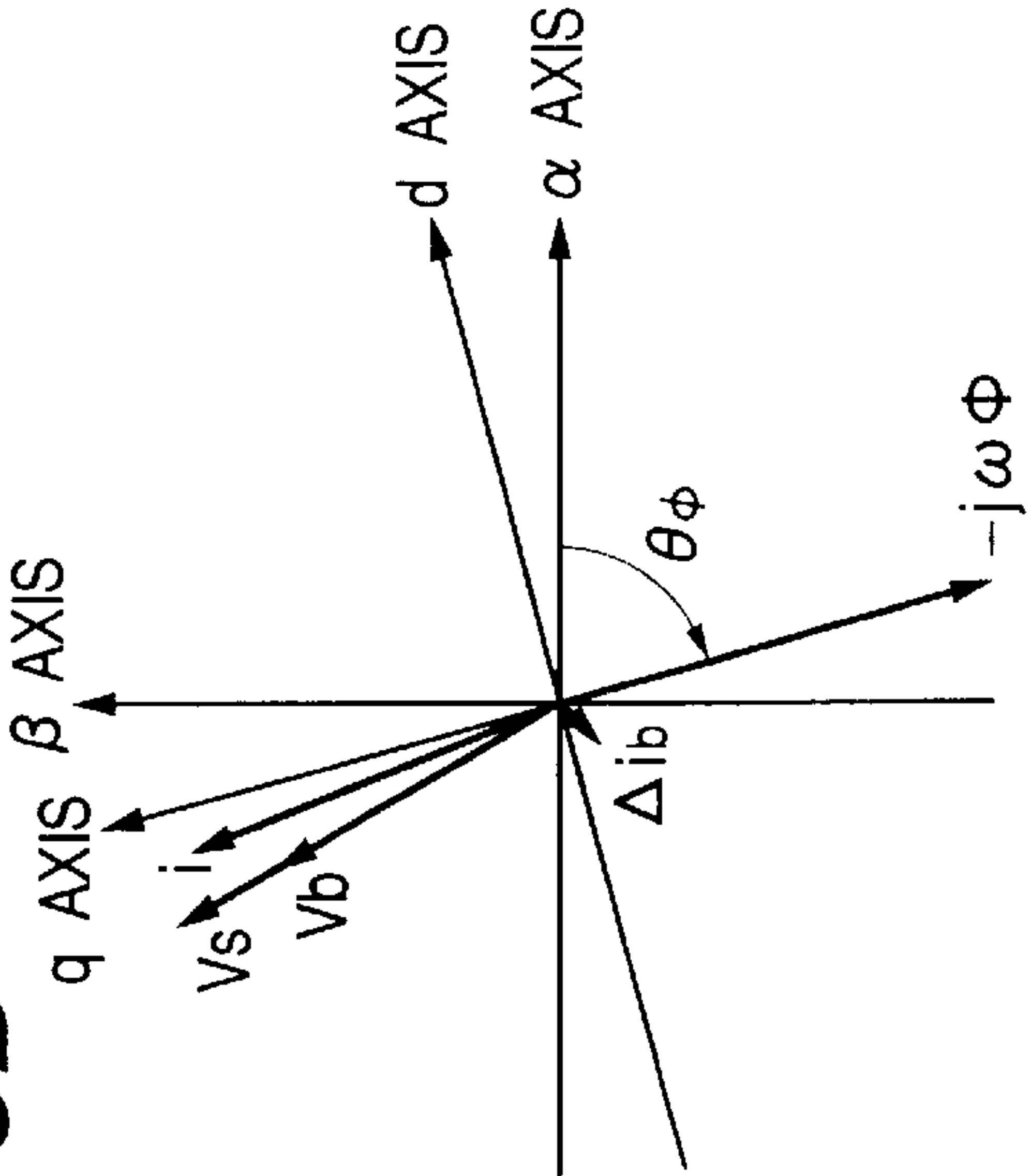


FIG. 13C

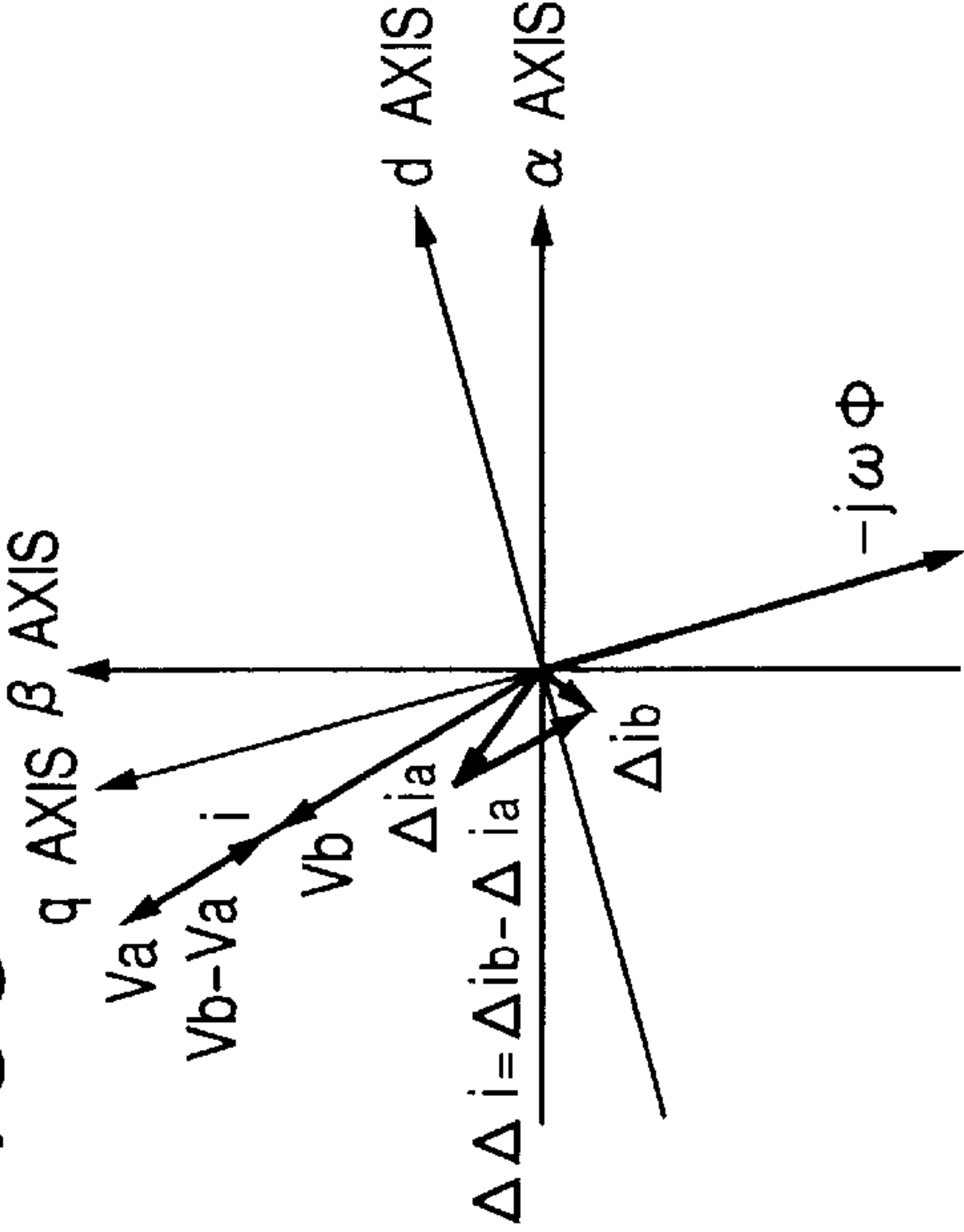


FIG. 13D

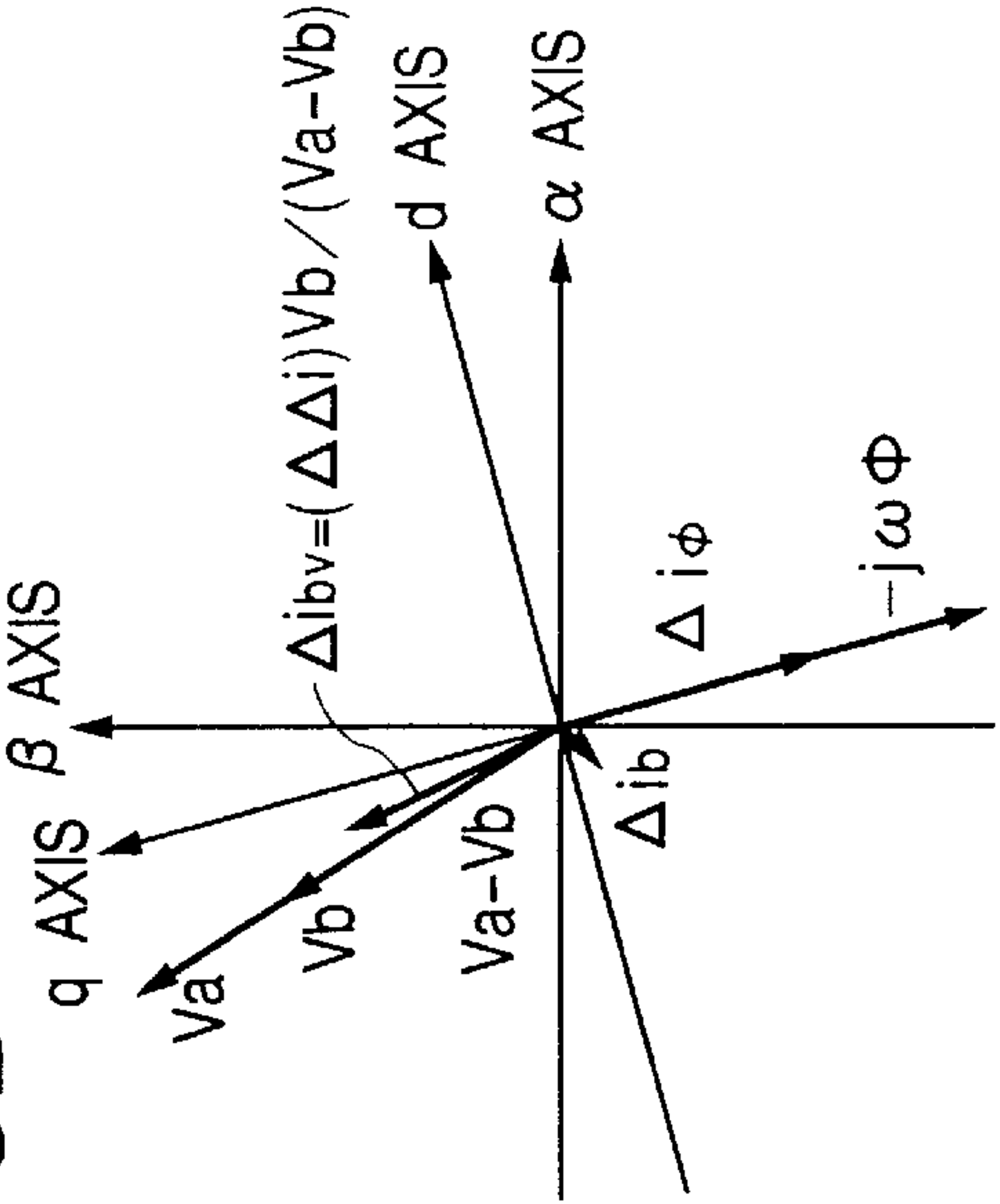


FIG. 14

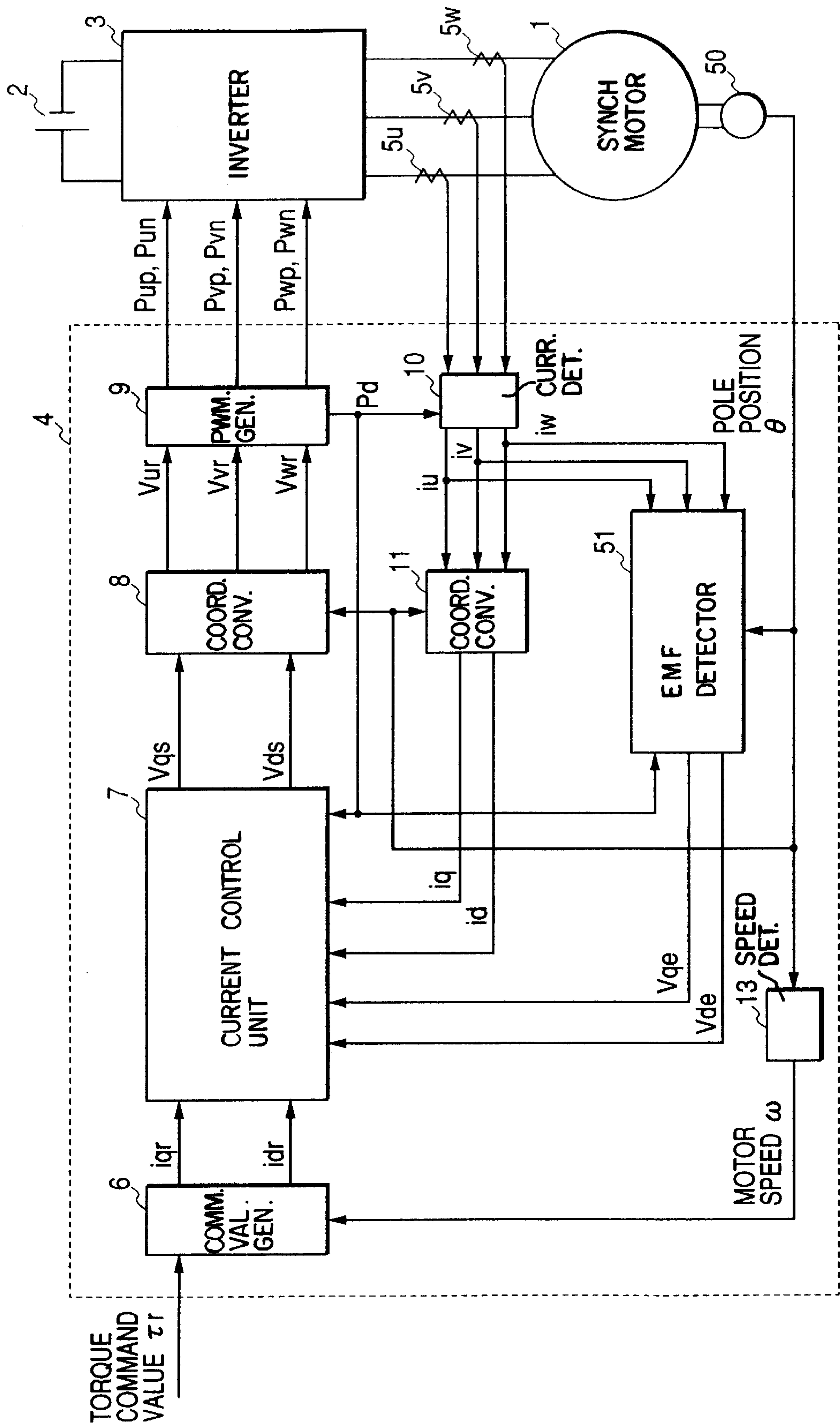


FIG. 15

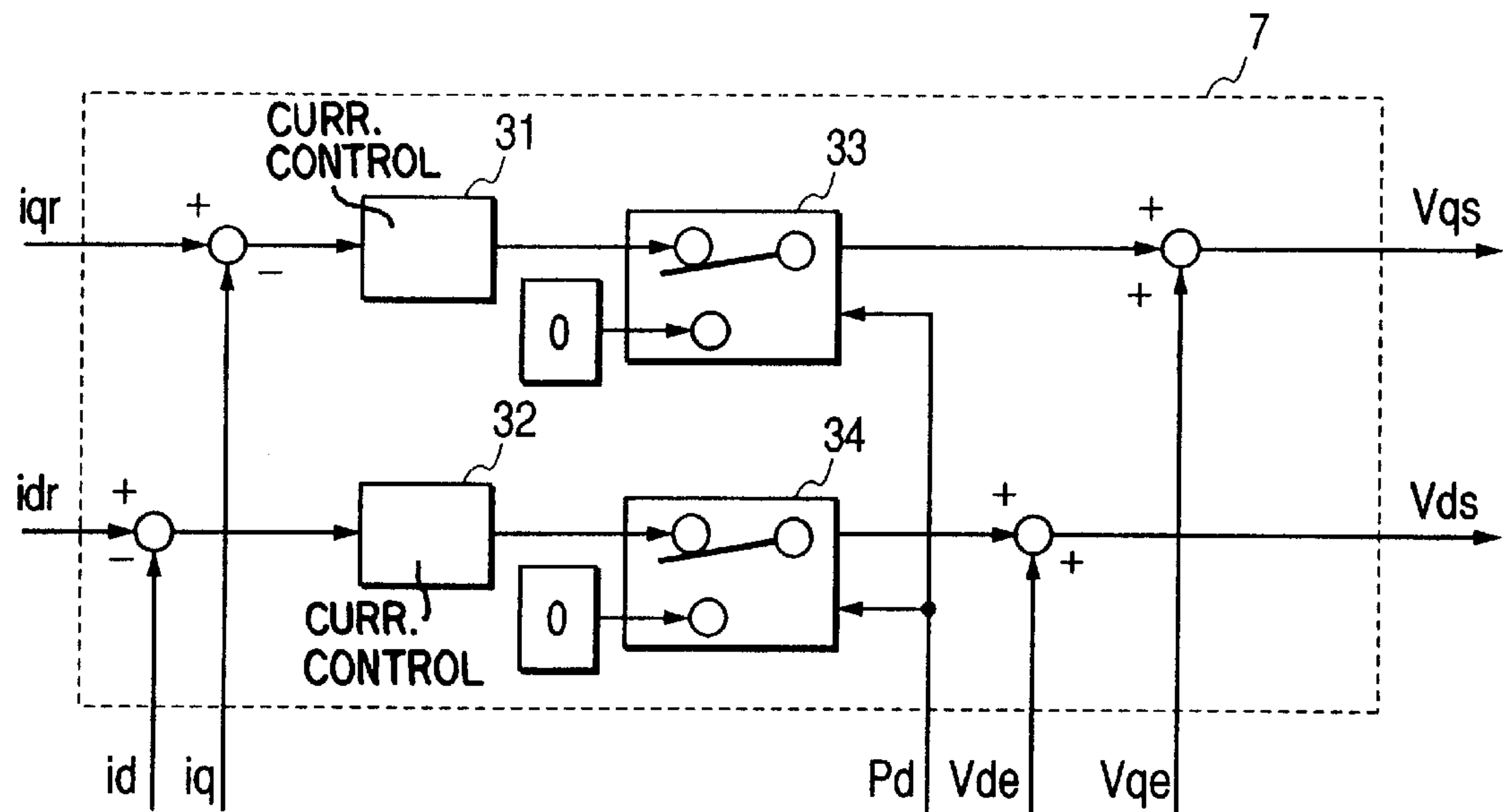


FIG. 16

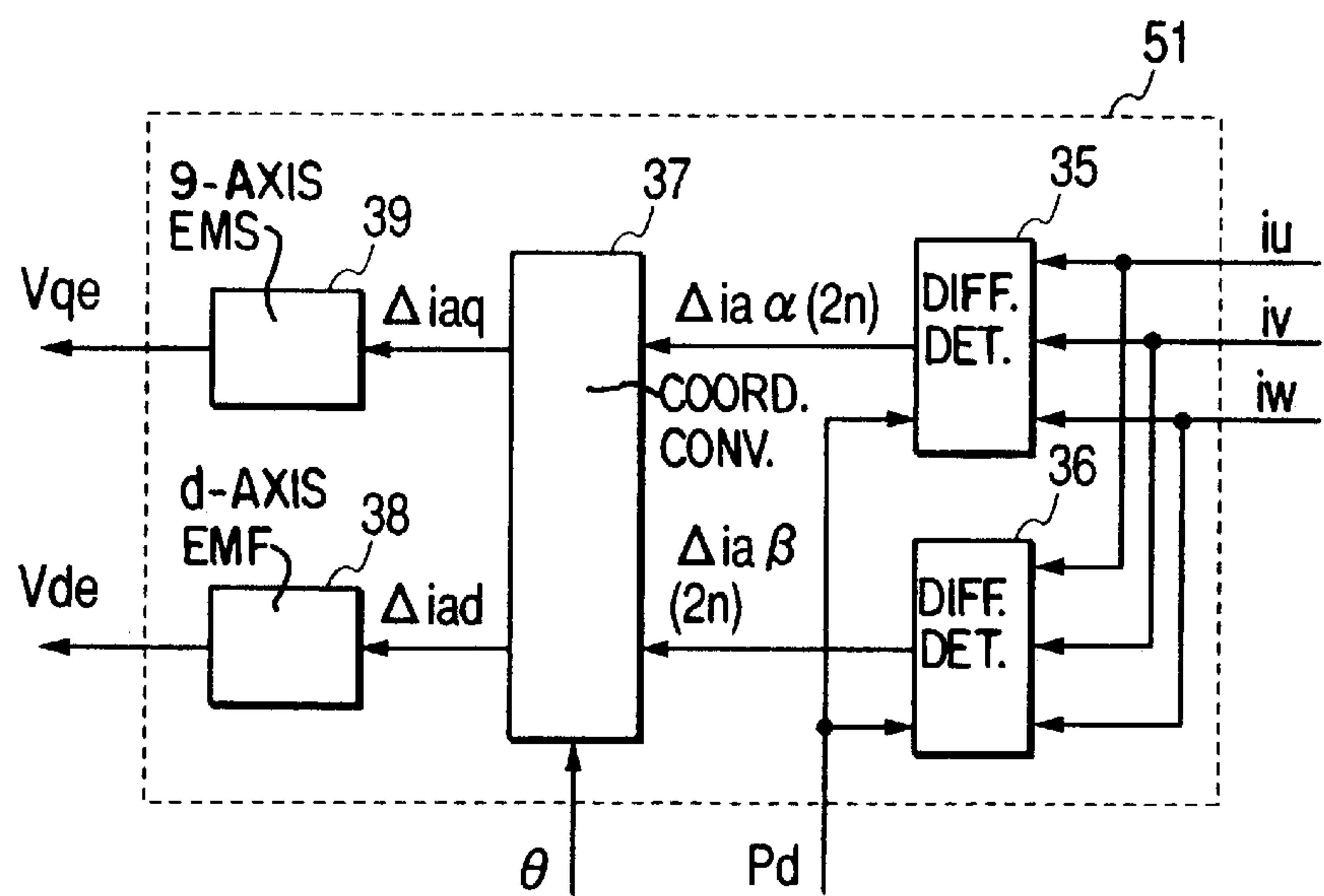


FIG. 17

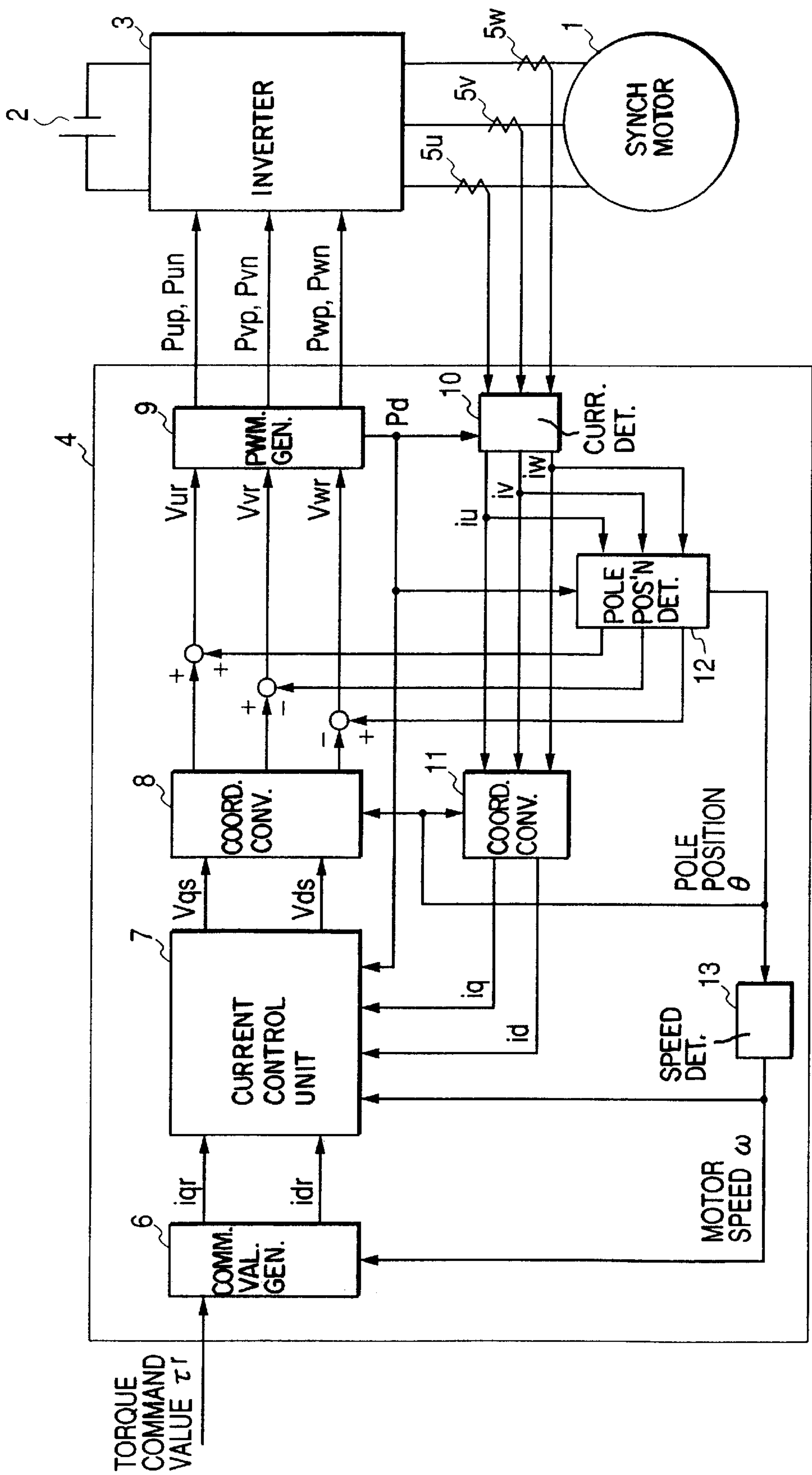


FIG. 18

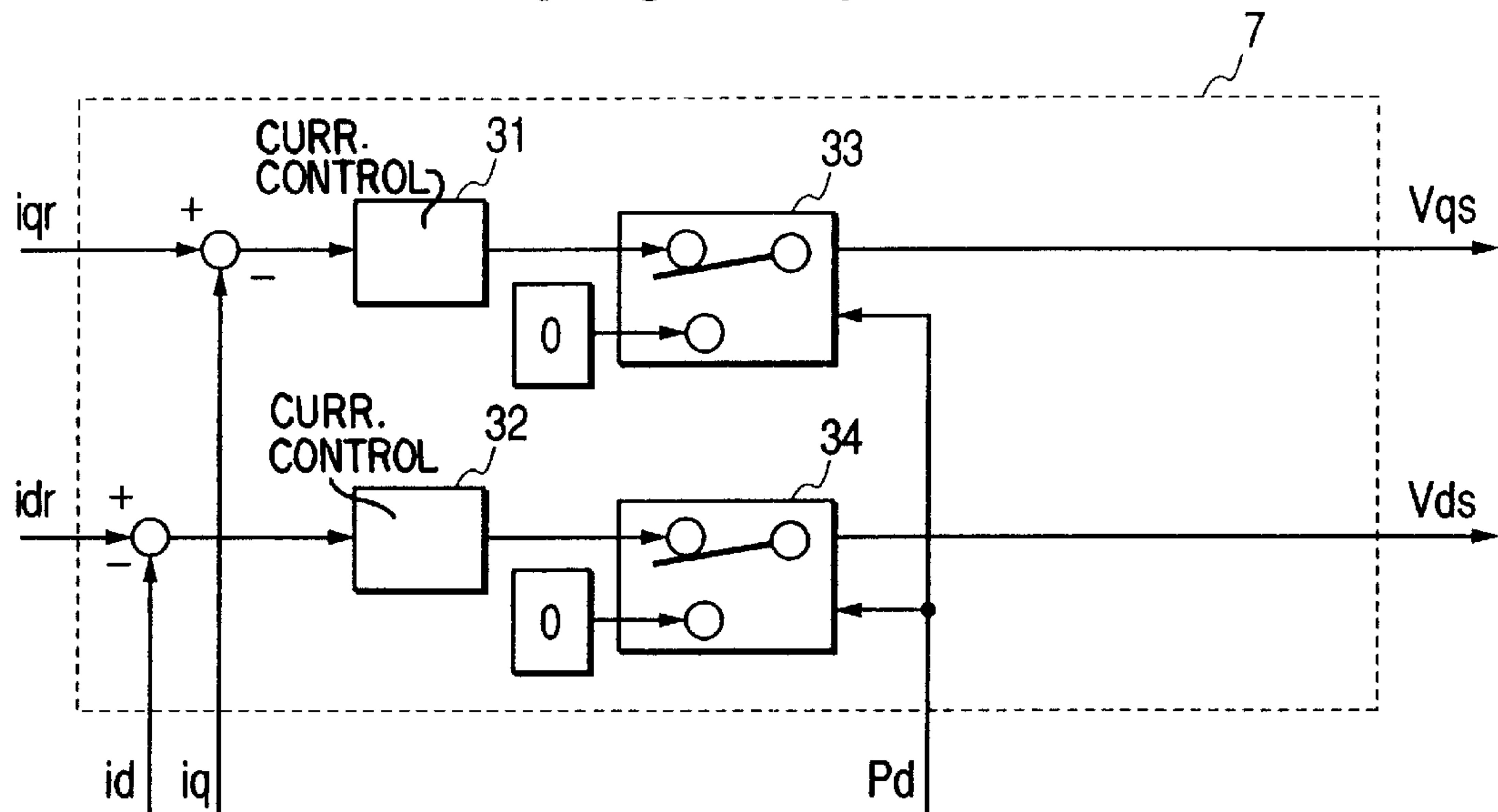
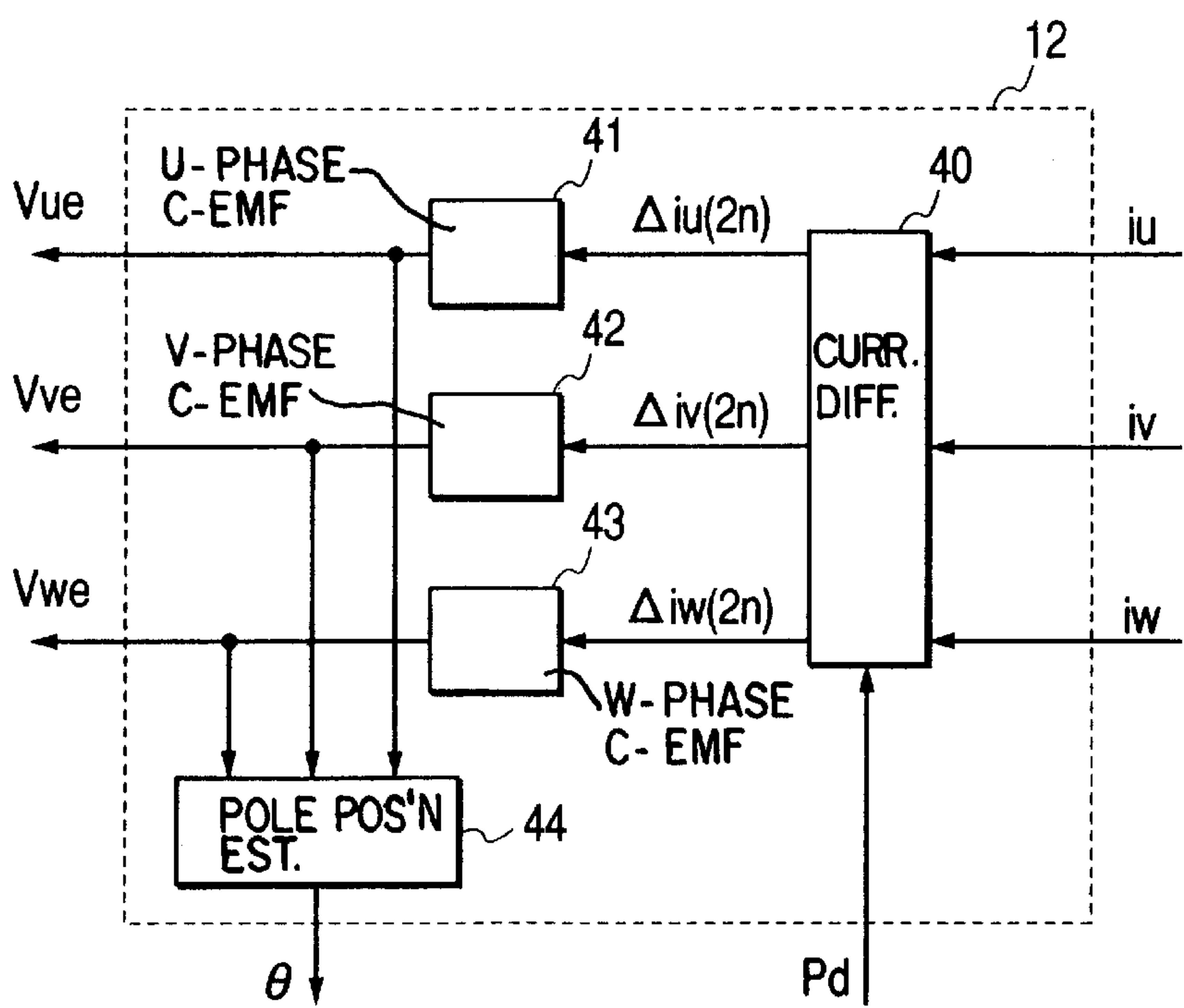


FIG. 19



MOTOR CONTROL DEVICE**BACKGROUND OF THE INVENTION****1. Field of the Invention**

The present invention relates to a motor control device which controls a synchronous motor including a reluctance motor with a high performance, and, in particular, relates to a motor control device which performs the control with a sensorless manner.

2. Conventional Art

In order to control such as speed and torque of a synchronous motor, it is necessary to detect or estimate its magnetic pole position. Thus, such as speed and torque of the synchronous motor can be controlled by performing a current control or a voltage control thereof based on the detected magnetic pole position.

Recently, methods of controlling a synchronous motor with no magnetic pole position sensor has been proposed.

A first method thereof is disclosed, for example, in JP-A-7-245981 (1995) and in a paper No. 170 at Heisei 8th National Meeting of Industrial Application Division for Japan Electrical Engineering's Society, which relates to a method of estimating the magnetic pole position based on a parallel motor current component and an orthogonal motor current component (current components in rotary coordinate system) in response to AC voltage application, and is characterized in that the detection of a magnetic pole position at a standstill and during a low speed rotation of the motor can be realized without using a magnetic pole position sensor.

A second method of superposing an additional voltage is disclosed, for example, in JP-A-11-150983 (1999) and JP-A-11-69884 (1999) in which method an additional voltage is applied so as to prevent magnetic flux saturation even in a high torque region, thereby, sensorless detection of a magnetic pole position is realized in a range from a low load to a heavy load at a standstill or during a low speed rotation.

Further, a third method is, for example, disclosed in JP-A-8-205578 (1996) in which a salient pole property of a synchronous motor is detected from a correlation between a voltage vector applied to the synchronous motor through a pulse width modulation control (PWM control) and a motor current ripple component (in vector amount) corresponding thereto. Further, since the third method utilizes usual PWM signals for controlling the synchronous motor, an advantage is obtained that no additional signals for the detection is required.

However, when detecting the magnetic pole position with the first method while driving the motor, it is necessary to extract a current having the same frequency component as that of an AC voltage used for the detection through such as a band pass filter using a notch filter and a Fourier integration. In particular, when the motor rpm increases, separation between motor input frequencies and AC voltage frequencies used for the detection becomes difficult, thus, a problem is posed that a stable motor drive at a high rpm is difficult. Further, the method requires a measure so as not to be affected by the switching characteristic of the inverter concerned. Namely, in contrast to the carrier wave frequencies of PWM signals which are from several kHz to 20 kHz, the frequencies of the AC voltage used for the detection are low at about several 100 Hz, therefore, noises of several 100 Hz may be generated when the motor is driving.

The second method is intended to improve performance when the motor is driven from a standstill condition or at a

low speed rotating condition, however, the timing of current detection and the relation with the PWM signals which become important for a high speed drive of the motor are not disclosed as well as no measures for realizing the position detection with a high accuracy are disclosed.

Further, in order to realize the third method it is necessary to detect correlation between the motor current condition and the applied voltage every time when the PWM signals vary. Namely, it is the base requirement that the detection of the motor current condition and recognition of the applied voltage condition have to be performed at least six times for every one cycle of the carrier waves which requires a high performance controller.

SUMMARY OF THE INVENTION

A first object of the present invention is to provide a motor control device which estimates a magnetic detection of motor current over a broad range from a standstill condition to a high speed rotating condition while using a non-expensive controller and controls a synchronous motor including a reluctance motor with a high response characteristic.

A second object of the present invention is to provide a high performance motor control device which can compensate a current follow-up property based on counter electromotive force information even if the motor speed varies.

According to the present invention, a motor control device is provided with an AC motor, an electric power inverter which applies a voltage to the AC motor and a control unit which controls the applied voltage with PWM signals in synchronism with carrier waves, wherein a magnetic pole position on a rotor of the AC motor is estimated through detection of current in the AC motor in synchronism with the carrier waves. For example, in a synchronous motor having a salient characteristic a current in the motor is detected while varying the applied voltage for every half cycle of the carrier waves and a current difference vector (a vector in static coordinate system) for the every half cycle is determined. Subsequently difference of twice current difference vectors (hereinbelow called as current difference difference vector) and difference corresponding to twice applied voltage vectors (hereinbelow called as voltage difference vector) are calculated, and the applied voltage (voltage difference vector) is controlled so that the phase difference thereof assumes 0. When the phase difference is rendered 0 with the above method, the phase of the voltage difference vector assumes the direction (d-axis) of the magnetic pole position, thereby, a magnetic pole position sensorless control can be realized. Since the change of the applied voltage is performed for every half cycle of the carrier waves, only the phase of the PWM signals is shifted which prevents generation of noises. Further, since the current is detected in synchronism with the carrier waves, the average values of the applied voltages for respective phases are accurately recognized which provides a characteristic that the relation between the applied voltage and the current difference can be detected in a short time represented by the carrier wave cycle. For this reason, the present invention is effective for sensorless approach in a field of motor control requiring a high response property.

Further, through the use of information on the current difference vector the counter electromotive force of the synchronous motor can be detected accurately. By making use of the thus detected counter electromotive force a counter electromotive force compensation in the current control system is effected, thereby, a control system having

a good current follow-up property is constituted in comparison with the conventional compensation of counter electromotive force which was estimated from the speed.

Through the detection of current in synchronism with the carrier wave cycle, the inventors found out a variety of methods of detecting the magnetic pole position other than the above explained which will be explained below in detail.

BRIEF DESCRIPTION OF THE DRAWINGS

FIG. 1 is a system diagram showing one embodiment to which the present invention is applied which detects a magnetic pole position of a synchronous motor by detecting current variation in the synchronous motor in synchronism with carrier waves for generating PWM signals;

FIG. 2 is a time chart showing a relationship between carrier waves when generating conventional PWM signals and the PWM signals;

FIG. 3 is a time chart showing a relationship between three phase voltage command values, carrier wave signals and PWM signals for detecting current difference values under three phase short circuit condition, and a timing of current sampling according to the present invention;

FIG. 4 is a flowchart showing a calculation method for detecting a magnetic pole position in FIG. 1 embodiment;

FIG. 5 is a control block diagram showing a modification of a magnetic pole position detection unit 12 in FIG. 1 embodiment;

FIGS. 6A through 6D are current and voltage vector diagrams of a synchronous motor for explaining the principle of magnetic pole position detection in FIG. 5 modification;

FIG. 7 is a control block diagram showing another modification of a magnetic pole position detection unit 12 in FIG. 1 embodiment which permits selection of a proper calculation method depending on a motor speed;

FIG. 8 is a flowchart showing processing contents performed in a second magnetic pole position estimation unit 20 in FIG. 7 modification;

FIGS. 9A through 9C are current and voltage vector diagrams of a synchronous motor explaining a principle of magnetic pole position detection performed in the second magnetic pole position estimation unit 20 as shown in FIG. 7;

FIG. 10 is a block diagram of a magnetic pole position sensorless control system for a synchronous motor in which calculation method of magnetic pole position detection is simplified by varying d-axis voltage command;

FIG. 11 is a block diagram of a magnetic pole position sensorless control system for a synchronous motor in which a magnetic pole position is estimated by detecting a counter electromotive force while varying the magnitude of voltage command value under in-phase condition;

FIG. 12 is a flowchart for explaining processings performed in a magnetic pole position estimation unit 30 in FIG. 11 embodiment;

FIGS. 13A through 13D are voltage and current difference vector diagrams for a synchronous motor for explaining a principle performed in FIG. 11 embodiment;

FIG. 14 is a control block diagram representing still another embodiment according to the present embodiment in which a characteristic of current control system is improved by detecting a counter electromotive force accurately;

FIG. 15 is a block diagram showing processing contents performed in a current control unit 7 in FIG. 14 embodiment;

FIG. 16 is a block diagram showing processing contents performed in a counter electromotive force detection unit 51 in FIG. 15 embodiment;

FIG. 17 is a block diagram of a motor control system representing a further embodiment according to the present invention in which a characteristic at a moment of motor speed sudden change is improved better than a conventional current control without using a magnetic pole position sensor;

FIG. 18 is a block diagram of a current control unit 7 in FIG. 17 embodiment for explaining a calculation method performed therein; and

FIG. 19 is a block diagram of an embodiment of the magnetic pole position detector in FIG. 17.

DETAILED DESCRIPTION OF THE EMBODIMENTS

Hereinbelow, an embodiment of the present invention will be explained with reference to FIG. 1. The present embodiment relates to a method in which a magnetic pole position is detected by making use of a counter electromotive force of a synchronous motor 1.

FIG. 1 is a block diagram of a motor control system in which the synchronous motor 1 is driven with DC energy from a battery 2. A DC voltage of the battery 2 is inverted by an inverter 3 into three phase AC voltage which is applied to the synchronous motor 1. The applied voltage is determined by a controller 4 after performing the following calculation. At first, in a current command value generating unit 6 d-axis current command value i_{dr} and q-axis current command value i_{qr} are determined with respect to a torque command value τ_r to be generated by the motor 1. Herein, d-axis designates the direction of magnetic pole position (magnetic fluxes) and q-axis designates the direction electrically orthogonal to the d-axis and with both of which d-q axis coordinate system is constituted. In the synchronous motor 1, the ratio of i_{dr} and i_{qr} can be varied under the condition of the same motor speed ω and the same motor torque τ being generated, although motor loss thereof varies. Therefore, when a motor speed ω is inputted to the current command value generating unit 6, optimum i_{dr} and i_{qr} which show a minimum motor loss with respect to a torque command value τ_r are designed to be outputted. Further, the motor speed ω is detected in a speed detection unit 13 based on a variation amount of magnetic pole position θ which will be explained later.

When a rotor with a magnet rotates, the d-q axis coordinate system also rotates, therefore, the phase from the static coordinate system (α - β axis coordinate system) is assumed as θ . Namely, the object of the present embodiment is to detect the phase θ of the magnetic pole (hereinbelow called as magnetic pole position θ) from a current. If d-axis current and q-axis current can be controlled in an alignment with the command values, the synchronous motor 1 can generate a torque to meet with the command value τ_r . Further, the torque command value τ_r is either commanded directly as illustrated or may be commanded from a speed control calculation circuit (not shown).

Further, U phase current i_u and V phase current i_v detected by current sensors 5u and 5v are sampled in a current detection unit 10 at the timing of current detection pulses pd which will be explained later, and are converted into d-axis current i_d and q-axis current i_q in d-q axis coordinate system at a coordinate conversion unit 11. In the present embodiment, currents detected in the current detection unit 10 are two phase currents i_u and i_v for U and V

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phases, because current i_w can be determined from i_u and i_v , the detection of W phase current i_w is omitted. However, the present invention is of course applicable when all of the three phase currents are detected. In a current control unit 7, d-axis current deviation between d-axis current command value i_{dr} and d-axis current i_d and q-axis current deviation between q-axis current command value i_{qr} and q-axis current i_q are calculated, and d-axis voltage command value V_{ds} and q-axis voltage command value V_{qs} for the respective current deviation are obtained through proportion and integration control calculation. Further, as a control method of compensating counter electromotive force a method of performing a non-interfering control making use of motor speed ω has also been proposed. In a coordinate conversion unit 8 to which d-axis voltage command value V_{ds} and q-axis voltage command value V_{qs} are inputted, three phase voltage command value V_{us} , V_{vs} and V_{ws} for the static coordinate system are calculated according to magnetic pole position θ . The three phase voltage command values are respectively added to detection use voltage command values V_{up} , V_{vp} and V_{wp} outputted from a magnetic pole position detection unit 12 which will be explained later, and are inputted to a PWM signal generating unit 9. Through calculation in the PWM signal generating unit 9 three phase PWM pulses P_{up} , P_{vp} , P_{wp} , P_{un} , P_{vn} and P_{wn} are outputted to the inverter 3. Thereby, a voltage to be applied to the synchronous motor 1 is determined.

Processing contents of the PWM signal generating unit 9 when the detection use voltage command values are $V_{up}=V_{vp}=V_{wp}=0$ will be explained. Through comparison of the waveforms of the respective phase voltage command values V_{ur} , V_{vr} and V_{wr} with saw teeth carrier waves, the three phase PWM pulses P_{up} , P_{vp} and P_{wp} can be obtained. Further, when a short circuiting prevention period is neglected, the PWM pulses P_{un} , P_{vn} and P_{wn} are respectively inversions of P_{up} , P_{vp} and P_{wp} , therefore, an explanation on the PWM pulses P_{un} , P_{vn} and P_{wn} is omitted. When duty of a PWM pulse exceeds over 50%, the average output voltage thereof assumes 0 or a positive value and when the duty is below 50%, the average output voltage assumes a negative value. In FIG. 2, the voltage command values V_{ur} , V_{vr} and V_{wr} are renewed at the time when the carrier waves assume their maximum values (for example, at time $t(2n)$, $t(2n+2)$). The PWM pulses P_{up} , P_{vp} and P_{wp} between times $t(2n)$ and $t(2n+2)$ show respectively symmetric shapes with respect to time $t(2n+1)$, therefore, the average values of the applied voltages for the respective phase from time $t(2n)$ to time $t(2n+2)$ are equal to those from time $t(2n+1)$ to time $t(2n+2)$.

Now the magnetic pole position detection unit 12 which is one of important elements in FIG. 1 embodiment will be explained. The magnetic pole position detection unit 12 is inputted of the U phase current i_u and the V phase current i_v and output the detection use voltage command values V_{up} , V_{vp} and V_{wp} according to flowchart as shown in FIG. 4 as well as calculates the magnetic pole position θ based on i_u and i_v . A processing method performed in the magnetic pole position detection unit 12 will be explained with reference to FIG. 4. At step 101, as detection use voltage command values $V_{up}(2n)$, $V_{vp}(2n)$ and $V_{wp}(2n)$ from time $t(2n)$ to time $t(2n+1)$, $-V_{us}(2n)$, $-V_{vs}(2n)$ and $-V_{ws}(2n)$ are outputted (wherein n is an integer). It is designed that at time $t(2n)$ these values are automatically set and the voltage command values V_{ur} , V_{vr} and V_{wr} for respective phases from time $t(2n)$ to time $t(2n+1)$ assume respectively 0. More specifically, the voltage command values in time periods $t(2n)\sim t(2n+1)$ and $t(2n+2)\sim t(2n+3)$ in the time chart as

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shown in FIG. 3 assume 0. Subsequently, at step 103 currents $i_u(2n)$ and $i_v(2n)$ at time $t(2n)$ are inputted. In FIG. 3, in synchronism with the leading edge of current detection pulses P_d which are generated at times $t(2n)$ and $t(2n+2)$, the detected currents are inputted. At step 104, as the detection use voltage command values $V_{up}(2n+1)$, $V_{vp}(2n+1)$ and $V_{wp}(2n+1)$ from time $t(2n+1)$ to time $t(2n+2)$, $V_{us}(2n)$, $V_{vs}(2n)$ and $V_{ws}(2n)$ are respectively outputted. In FIG. 3, it is implied that the voltage command values in the time period $t(2n+1)\sim t(2n+2)$ are respectively $2V_{us}(2n)$, $2V_{vs}(2n)$ and $2V_{ws}(2n)$. When performing these controls, average values of the voltage command values in the time period $t(2n)\sim t(2n+2)$ assume $V_{us}(2n)$, $V_{vs}(2n)$ and $V_{ws}(2n)$, and it is obtained that although the phases of the PWM pulses P_{up} , P_{vp} and P_{wp} vary from those in FIG. 2, but the pulse widths thereof are equal to those in FIG. 2. Namely, it is equivalent that a control of shifting phases of the PWM pulses is performed. Although the detection use voltage command values are added, the present embodiment shows a characteristic that no substantial influence is affected to the current control performance. Further, there was a problem conventionally that when detection use AC voltage is applied, noises are generated, however, since the detection use voltage command values are controlled for every half cycle of the carrier waves in synchronism therewith, the present embodiment shows a characteristic that no noises exceeding those caused by the normal PWM switching are generated.

At step 106, currents $i_u(2n+1)$ and $i_v(2n+1)$ at time $t(2n+1)$ are inputted. Subsequently, at step 107 variation amounts of respective phase currents $i_u(2n+1)\sim i_u(2n)$, $i_v(2n+1)\sim i_v(2n)$, namely current difference values $\Delta i_u(2n)$ and $\Delta i_v(2n)$ from time $t(2n)$ to time $t(2n+1)$ are determined. At the subsequent step 108, phase θ_q of the current difference vector $\Delta i(2n)$ is obtained from the current difference values $\Delta i_u(2n)$ and $\Delta i_v(2n)$. The applied voltages at time period $t(2n)\sim t(2n+1)$ are the same for the respective phases, in that assume zero voltage vector state. Herein, it is assumed that a current difference vector $\Delta i(2n)$ of the synchronous motor 1 varies by its counter electromotive force, the phase of the current difference vector $\Delta i(2n)$ assumes in negative direction in q-axis. Accordingly, the phase advanced by $\pi/2$ [rad] with respect to the phase θ_q of the current difference vector Δi corresponds to the magnetic pole position, therefore, at step 109, the magnetic pole position θ is determined by adding $\pi/2$ [rad] to the phase θ_q . Namely, through the control of shifting the respective PWM pulses the magnetic pole position detection can be realized only with the current changes in the half cycle of the carrier waves, therefore, the advantages that the magnetic pole position is rapidly detected without increasing noises. Further, it is unnecessary to use motor parameters including those of a synchronous motor having a salient pole characteristic, therefore, a further advantage that the magnetic pole position can be detected further accurately. Still further, when it is required to take into account of a dead time, the applied voltage can be adjusted so to assume zero voltage vector depending on the current directions of the respective phases.

FIG. 5 is a modification of the magnetic pole position detection unit 12 which is effective when a synchronous motor 1 has an inverted salient pole characteristic (which implies that d-axis inductance is smaller than q-axis inductance). The magnetic pole position detection unit 12 as shown in FIG. 5 is constituted by a current difference calculation unit 14, a current difference difference vector phase calculation unit 15, a magnetic pole position estima-

tion unit **16** and a detection use voltage generation unit **17**. At first, in the detection use voltage generation unit **17** the detection use voltage command values $V_{up}(2n)$, $V_{vp}(2n)$ and $V_{wp}(2n)$ from time $t(2n)$ to time $t(2n+1)$ are calculated according to the following equations (1)-(3) and the calculation results are outputted therefrom;

$$V_{up}(2n) = V_0 \cos(\theta_{vp}(2n)) \quad (1)$$

$$V_{vp}(2n) = V_0 \cos(\theta_{vp}(2n) + 2\pi/3) \quad (2)$$

$$V_{wp}(2n) = V_0 \cos(\theta_{vp}(2n) + 4\pi/3) \quad (3)$$

wherein, $\theta_{vp}(2n)$ is an output at time $t(2n)$ from the magnetic pole position estimation unit **16**. Accordingly, the vector phase of the detection use voltage command value is $\theta_{vp}(2n)$. Subsequently, the detection use voltage command values $V_{up}(2n+1)$, $V_{vp}(2n+1)$ and $V_{wp}(2n+1)$ from time $t(2n+1)$ to time $t(2n+2)$ are calculated according to the following equations (4)-(6) and the calculation results are outputted from the detection use voltage generation unit **17**;

$$V_{up}(2n+1) = V_0 \cos(\theta_{vp}(2n) + \pi) \quad (4)$$

$$V_{vp}(2n+1) = V_0 \cos(\theta_{vp}(2n) + 5\pi/3) \quad (5)$$

$$V_{wp}(2n+1) = V_0 \cos(\theta_{vp}(2n) + \pi/3) \quad (6)$$

The vector phase of the detection use voltage command value at this instance is $\theta_{vp}(2n) + \pi$, namely directs to the opposite direction as the vector phase of the detection use voltage command value from time $t(2n)$ to time $t(2n+1)$.

In the current difference calculation unit **14**, current difference values $\Delta i_u(2n)$ and $\Delta i_v(2n)$ from time $t(2n)$ to time $t(2n+1)$ are respectively calculated from the currents $i_u(2n)$ and $i_v(2n)$ at time $t(2n)$ and the currents $i_u(2n+1)$ and $i_v(2n+1)$ at time $t(2n+1)$, and a current difference vector $\Delta i_a(2n)$ is determined according to these calculated values. Further, current difference values $\Delta i_u(2n+1)$ and $\Delta i_v(2n+1)$ from time $t(2n+1)$ to time $t(2n+2)$ are calculated from currents $i_u(2n+1)$ and $i_v(2n+1)$ at time $t(2n+1)$ and currents $i_u(2n+2)$ and $i_v(2n+2)$ at time $t(2n+2)$ and the current difference vector $\Delta i_b(2n+1)$ is determined according to these calculated values. The difference between the current difference vector $\Delta i_b(2n+1)$ and the current difference vector $\Delta i_a(2n)$ is the current difference difference vector $\Delta \Delta i(2n)$, and its phase $\theta_{ip}(2n)$ is calculated in the current difference vector phase calculation unit **15**. It is to be noted that both the current difference vector and the current difference difference vector are values in the static coordinate system of which importance will be explained later.

The magnetic pole position estimation unit **16** performs a control so as to make the vector phase $\theta_{vp}(2n)$ of the detection use voltage command value come close to the phase $\theta_{ip}(2n)$ of the current difference difference vector $\Delta \Delta i(2n)$. Through this control when the phase θ_{ip} stably coincides with the phase θ_{vp} , the phase θ_{vp} assumes d-axis, namely corresponds to the magnetic pole position θ . Further, if the phase of d-axis varies more than several times during one cycle of the carrier waves (namely, in a case that the motor speed ω is high), a control of correcting the phase depending on motor speed ω is performed.

The control performed in FIG. 5 modification will be explained with reference to vector diagrams as shown in FIGS. 6A through 6D. FIG. 6A shows the state of current difference vector when only the three phase voltage command values V_{us} , V_{vs} and V_{ws} outputted from the coordinate conversion unit **8** are applied to the synchronous motor **1**, in that the voltage vector $V_s(2n)$ from time $t(2n)$ to time

$t(2n+2)$ corresponds to the three phase voltage command values V_{us} , V_{vs} and V_{ws} . The current difference vector $\Delta i(2n)$ is a vector sum of a first current difference vector $\Delta i_{vs}(2n)$, a second current difference vector $\Delta i_{\phi}(2n)$ and a third current difference vector $\Delta i_{ii}(2n)$ of which component are respectively affected by the voltage vector $V_s(2n)$, the counter electromotive force vector $-j\omega\phi(2n)$ and the current vector $i(2n)$. The second current differential vector $\Delta i_{\phi}(2n)$ assumes the same phase as the counter electromotive force vector $-j\omega\phi(2n)$ regardless to whether or not the synchronous motor **1** shows a salient pole characteristic (including a reverse salient pole characteristic). Further, when the interval from time $t(2n)$ to time $t(2n+1)$ is short in comparison with the time constant of the circuit in the synchronous motor **1**, the third current difference vector $\Delta i_{ii}(2n)$ is small in comparison with other current difference vectors and can be assumed as 0. When the phase of the voltage vector $V_s(2n)$ is neither directed to d-axis direction nor to q-axis direction, the phase of the first current differential vector $\Delta i_{vs}(2n)$ does not coincide with the phase of the voltage vector $V_s(2n)$ due to the salient pole characteristic of the synchronous motor **1**. As has been explained hitherto, since the current difference vector $\Delta i(2n)$ is affected by the voltage vector $V_s(2n)$ and the counter electromotive force vector $-j\omega\phi(2n)$, it is difficult to detect the magnetic pole position under this condition.

Therefore, during the time from $t(2n)$ to $t(2n+1)$ $V_a(2n)$ representing a vector sum of the voltage vector $V_s(2n)$ and the detection use voltage vector $V_p(2n)$ is applied as shown in the vector diagram in FIG. 6B. The detected current difference vector $\Delta i_a(2n)$ represents the vector sum of the current difference vector $\Delta i_{vp}(2n)$ induced by the detection use voltage vector $V_p(2n)$ and the current difference vector $\Delta i(2n)$ as shown in FIG. 6A. During the following interval from time $t(2n+1)$ to time $t(2n+2)$ $V_b(2n+1)$ representing a vector sum of the voltage vector $V_s(2n)$ and the detection use voltage vector $V_p(2n+1)$ is applied as shown in FIG. 6C. Herein as the detection use voltage vector $V_p(2n+1)$ a vector satisfying the following equation is provided;

$$V_p(2n+1) = -V_p(2n) \quad (7)$$

In this instance, the detected current difference vector $\Delta i_b(2n+1)$ represents a vector sum of the current difference vector $\Delta i_{vp}(2n+1)$ induced by the detection use voltage vector $V_p(2n+1)$ and the current difference vector $\Delta i(2n+1)$ as shown in FIG. 6A. When magnetic fluxes of the synchronous motor **1** are not saturated, $\Delta i_{vp}(2n+1)$ satisfies the following relationship;

$$\Delta i_{vp}(2n+1) = -\Delta i_{vp}(2n) \quad (8)$$

Further, during the interval from time $t(2n)$ to time $t(2n+2)$ when the motor speed is in a range that the rotation phase of the magnetic pole of the synchronous motor **1** can be treated small, $\Delta i(2n+1)$ in FIG. 6C substantially coincides with $\Delta i(2n)$ in FIG. 6B.

With reference to FIG. 6D, detection of the current difference vector $\Delta i_a(2n)$ from time $t(2n)$ to time $t(2n+1)$ and the current difference vector $\Delta i_b(2n+1)$ from time $t(2n+1)$ to time $t(2n+2)$ and the purpose of obtaining the current difference difference vector $\Delta \Delta i(2n)$ representing the difference therebetween will be explained. In view of the relationships shown in FIGS. 6B and 6C, it is understood that the difference between $\Delta i_a(2n)$ and $\Delta i_b(2n+1)$, namely the current difference difference vector $\Delta \Delta i(2n)$ assumes $2\Delta i_{vp}(2n)$. As has been explained previously, $\Delta i_{vp}(2n)$ is a component induced by applying the detection use voltage vector

$V_p(2n)$. Herein, it is assumed that the phases of $V_p(2n)$ and $\Delta i_p(2n)$ are respectively $\theta_{vp}(2n)$ and $\theta_{ip}(2n)$. For a synchronous motor 1 having an inverted salient pole characteristic only when $\theta_{vp}(2n)$ coincides with either d-axis, $\theta_{vp}(2n)$ coincides with $\theta_{ip}(2n)$. When $\theta_{vp}(2n)$ does not coincide with $\theta_{ip}(2n)$, $\Delta \Delta i(2n)$ (herein, indicated as $2\Delta i_p$) is placed closer to d-axis than $V_p(2n)$ as shown in FIG. 6D. Therefore, as explained in connection with the magnetic pole position estimation unit 16 as shown in FIG. 5, when the control of shifting $\theta_{vp}(2n)$ close to $\theta_{ip}(2n)$ is performed, the $\theta_{vp}(2n)$ stabilizes at d-axis. Further, whether positive or negative direction in d-axis can be estimated with reference to the counter electromotive force direction. When performing the above explained control, a torque control or a speed control of the synchronous motor having an inverted salient pole characteristic can be performed from a standstill condition to a high speed rotation with no magnetic pole position sensor. Moreover, since the magnetic pole position can be detected within one cycle of the carrier waves, it is possible to obtain substantially the same degree of response characteristic as when a magnetic pole position sensor is installed. Further, with regard to a synchronous motor having a salient pole characteristic other than the inverted salient pole characteristic the sensorless control can be performed on the same principle.

FIG. 7 is another embodiment of the synchronous motor control which permits to select one of a plurality of methods of detecting a magnetic pole position with no magnetic pole position sensor depending on motor speed ω . In the present embodiment, the overall motor control structure is the same as FIG. 1 embodiment, however, the structure of the magnetic pole position detection unit 12 is modified from those in FIGS. 1 and 5.

FIG. 7 shows a block diagram of the magnetic pole position detection unit 12 which is constituted by a current difference calculation unit 14, a first magnetic pole position detection unit 19, a second magnetic pole position detection unit 20, a third magnetic pole position detection unit 21 and a calculation method change-over unit 18. The current difference calculation unit 14 performs the calculation as explained in FIG. 5 modification. Likely, in the first magnetic pole position detection unit 19 the calculation as explained in connection with FIG. 5 embodiment is performed. Namely, the principle of magnetic pole position estimation by this calculation makes use of the current variation in relation to the salient pole characteristic (or the inverted salient pole characteristic) of the synchronous motor 1, the magnetic pole position detection can be performed from the standstill condition to a medium speed region of the synchronous motor with this calculation method.

On the other hand, the second magnetic pole position detection unit 20 relates to a method of extracting the current difference vector due to the counter electromotive force from information on the applied voltage. The present method is effective for a motor speed range from a medium speed to a high speed.

Further, the third magnetic pole position detection unit 21 relates to an estimation method which makes use of a current difference vector caused by magnetic flux rotation without varying the applied voltage, the present method is particularly effective for magnetic pole position detection at the time of high speed rotation of the motor.

Now, at first the second magnetic pole position calculation unit 20 will be explained, which is constituted by an in-short circuit current difference phase calculation unit 22, a magnetic pole position estimation unit 23 and detection use

voltage generation unit 24. The in-short circuit current difference phase calculation unit 22 calculates the phase of the counter electromotive force and estimates the magnetic pole position through performing the calculation as shown in FIG. 8 flowchart. FIGS. 9A through 9C vector diagrams of current and voltage at this instance. At step 110 in FIG. 8, current difference vectors $\Delta i_a(2n)$ and $\Delta i_b(2n+1)$ are inputted. $\Delta i_a(2n)$ is the current difference vector, when a voltage $V_a(2n)$ is applied during interval $t(2n) \sim t(2n+1)$. Like the vector diagram as has been explained in connection with FIG. 6B, $V_a(2n)$ is a vector sum of the voltage vector $V_s(2n)$ and the detection use voltage vector $V_p(2n)$ of which relationship is shown in FIG. 9A. Likely, $\Delta i_b(2n+1)$ is a current difference vector, when voltage $V_b(2n)$ is applied during interval $t(2n+1) \sim t(2n+2)$. The applied voltage $V_b(2n+1)$ at this instance is a vector sum of the voltage vector $V_s(2n)$ and the detection use voltage vector $-V_p(2n)$ of which relationship is shown in FIG. 9B. At the subsequent step 111, a current difference vector $\Delta i_\phi(2n)$ due to a counter electromotive force is calculated from a vector coordinate defined by two crossing straight lines, one running in parallel with the applied voltage $V_a(2n)$ and passing on the apex of $\Delta i_a(2n)$ and the other running in parallel with the applied voltage $V_b(2n+1)$ and passing on the apex of $\Delta i_b(2n+1)$, of which implication will be explained with reference to FIG. 9C. When the inverted salient pole characteristic is small, a current difference component $\Delta i_{aq}(2n)$ among the current difference vector $\Delta i_a(2n)$ which is orthogonal to the applied voltage $V_a(2n)$ is hardly affected by the applied voltage $V_a(2n)$. Namely, $\Delta i_{aq}(2n)$ can be treated mostly as a component of the current difference vector $\Delta i_\phi(2n)$ due to the counter electromotive force. Likely, the current difference component $\Delta i_{bq}(2n+1)$ among the current difference vector $\Delta i_b(2n+1)$ which is orthogonal to the applied voltage $V_b(2n+1)$ is hardly affected by the applied voltage $V_b(2n+1)$, and can be treated mostly as a component of the current difference vector $\Delta i_\phi(2n)$ due to the counter electromotive force. In other words, a component of the current difference vector $\Delta i_\phi(2n)$ which is orthogonal to $V_a(2n)$ is $\Delta i_{aq}(2n)$, and another component of $\Delta i_\phi(2n)$ which is orthogonal to $V_b(2n+1)$ is $\Delta i_{bq}(2n+1)$. Accordingly, when performing the calculation at step 111, the current difference vector $\Delta i_\phi(2n)$ can be determined. At step 112, the phase θ_ϕ of $\Delta i_\phi(2n)$ is calculated, of which phase is the very phase of the counter electromotive force as has been explained above. Since the phase θ_ϕ of the vector $\Delta i_\phi(2n)$ corresponds to q-axis (in negative direction) as shown in FIG. 9C, the magnetic pole position θ_d can be obtained by adding $\pi/2$ [rad] to this phase θ_ϕ of which calculation is performed at step 113. Thus, the magnetic pole position can be determined based on the counter electromotive force. Further, when the salient pole characteristic of the synchronous motor 1 is large, the processings as shown in FIG. 8 are performed in view of the salient pole characteristic of which effect varies depending on the speed, thereby, a further accurate magnetic pole position detection can be realized. In response to the magnetic pole position θ_d outputted from the in-short circuit current difference phase calculation unit 22 the magnetic pole position estimation unit 23 feeds back the magnetic pole position θ_{vp} which was determined by the processings performed until the last time, thereby, a filtering with respect to noises is performed. In this instance, through performing positional compensation with respect to motor speed ω , varying magnetic pole position during one sampling period in which the processings are performed is taken into account. The detection use voltage generation unit 24 functions to output the detection

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use applied voltage V_p as shown in FIGS. 9A and 9B, and determines V_p according to the magnetic pole position θ_{vp} and d-axis voltage command value and q-axis voltage command value outputted from the current control unit 7. There are a variety of methods of determining V_p , however, in the present modification the calculation is performed so that V_p directs to the direction orthogonal to the voltage vector V_s as shown in FIGS. 9A through 9C, which is determined in a consideration that the absolute values of V_a and V_b do not exceedingly increase.

Further, the third magnetic pole position detection unit 21 is a means for performing the magnetic pole position detection with a high accuracy when the motor speed ω increases. The fundamental principle of the present invention is to detect a magnetic pole position from variation of current which is in synchronism with the carrier waves and since the time interval for the current variation used for the calculation is short (for example, one cycle time of the carrier waves), therefore, it is assumed that during the short time interval the variation of the magnetic pole position is very small. However, when the motor speed increases exceedingly high, it sometimes happens that the magnetic pole position varies more than 10° during the short time interval. Therefore, when the motor speed ω is such high, the magnetic pole position is calculated by detecting the variation of the current difference vector due to variation of the position (phase) of the counter magnetomotive force without varying the applied voltage of which method is effective when the motor speed ω is high, because it is unnecessary to increase the applied voltage.

The third magnetic pole position detection unit 21 is constituted by an under flux rotation current difference phase calculation unit 25, a magnetic pole position estimation unit 26 and a detection use voltage generation unit 27. The detection use voltage generation unit 27, in practice, always outputs value 0. The under flux rotation current difference phase calculation unit 25 determines a difference between the current difference vectors $\Delta i_a(2n)$ and $\Delta i_b(2n+1)$, namely the current difference difference vector $\Delta \Delta i(2n)$, of which calculation itself is identical to that performed in the current difference difference vector phase calculation unit 15. Since the detection use voltage V_{up} , V_{vp} and V_{wp} are always 0, the current difference difference vector $\Delta \Delta i(2n)$ will primarily be 0. However, an averaged magnetic pole position from time $t(2n)$ to time $t(2n+1)$ and an averaged magnetic pole position from time $t(2n+1)$ to time $t(2n+2)$ are different, therefore, average counter electromotive forces $V_{emf}(2n)$ and $V_{emf}(2n+1)$ at the two time intervals are different. Accordingly, due to a difference between $V_{emf}(2n+1)$ and $V_{emf}(2n)$, in that counter electromotive force difference vector $\Delta V_{emf}(2n)$ the current difference difference vector $\Delta \Delta i(2n)$ can not assume 0, but shows a certain amount of vector. Herein, the direction of $\Delta V_{emf}(2n)$ is almost orthogonal to the direction of q-axis, in that in the direction of d-axis (in negative direction). Accordingly, the direction of $\Delta \Delta i(2n)$ also assumes the direction of $\Delta V_{emf}(2n)$, in that d-axis direction (in negative direction thereof) $-\theta_d$. Through the calculation of the direction of the current difference difference vector $\Delta \Delta i(2n)$ in this way, $-\theta_d$ can be obtained. The magnetic position estimation unit 26 performs a noise processing by making use of the magnetic pole position θ and $-\theta_d$ obtained by the last processings, and outputs a magnetic pole position θ . As has been explained, the third magnetic pole position detection unit 21 estimates the magnetic pole position by making use of the current difference vector induced by the rotation of magnetic fluxes without varying the applied voltage of which method is

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particularly effective for the detection of the magnetic pole position at the time of high speed rotation of the synchronous motor.

The calculation method change-over unit 18 estimates a motor speed ω based on the magnetic pole position θ (or θ_{vp}) obtained by the magnetic pole position estimation units and performs processing of selecting an optimum magnetic pole position estimation method in response to the estimated motor speed.

With the present embodiment, the magnetic pole position can always be detected in a moment with a high accuracy over a motor speed range from 0 to its maximum speed, thereby, a motor control system having a high response and high performance control property can be provided. Further, when a current sensor having an accurate current detection and less noise property is used, the magnetic pole position can be estimated in a moment for every one cycle of the carrier waves, while eliminating the filtering processing at the magnetic pole position estimation units 23 and 26. Further, in the combination of the magnetic pole position detection methods to be changed-over, a further magnetic pole position detection method which will be explained later can be included.

FIG. 10 is still another embodiment showing a magnetic pole position detection method in which the calculation therefor is simplified. Primary differences of FIG. 10 embodiment from 1 and FIG. 5 embodiments are that the two current sensors are increased to three, the output V_p of the detection use voltage generation unit 28 is added to the d-axis voltage command, and the processing method in the magnetic pole position detection unit 12 is simplified. As has been explained in connection with FIG. 1 embodiment, when the phase currents i_u and i_v are detected, the W phase current i_w can be determined from i_u and i_v , therefore, the detection of the W phase current i_w can be omitted, however, in the present embodiment the W phase current i_w is likely detected with a current sensor 5w. The current detection unit 10 samples the phase currents i_u , i_v and i_w in response to the current detection pulses P_d and performs an offset compensation so as to keep the relationship $i_u+i_v+i_w=0$. Thereby, offset errors in the current sensors can be compensated and a highly accurate magnetic pole position detection can be realized.

Now, the detection use voltage generation unit 28 representing features of the present embodiment will be explained. As has been explained in connection with FIG. 5 embodiment, the detection use applied voltage V_p is applied in positive and negative direction of an estimated d-axis in synchronism with the carrier waves. For this reason, in the detection use voltage generation unit 28 a process of adding the detection use applied voltage V_p of positive and negative values alternatively to the d-axis voltage V_d in response to the current detection pulses P_d .

The magnetic pole position detection unit 12 is constituted by the current difference calculation unit 14 and the current difference difference vector phase calculation unit 15 which have been explained in connection with FIG. 5 embodiment. In the current difference calculation unit 14, the current difference vector $\Delta i_a(2n)$ during time from $t(2n)$ to $t(2n-1)$ and the current difference vector $\Delta i_b(2n+1)$ during time from $t(2n+1)$ to $t(2n+2)$ are calculated respectively from the three phase currents i_u , i_v and i_w , and are outputted to the current difference vector phase calculation unit 15. Subsequently, in the current difference difference vector phase calculation unit 15, the current difference difference vector $\Delta \Delta i(2n)$ is determined from a difference between $\Delta i_b(2n+1)$ and $\Delta i_a(2n)$, and the phase $\theta_{ip}(2n)$ is

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calculated. The phase $\theta_{ip}(2n)$ of $\Delta\Delta i(2n)$, as it is, is assumed as the magnetic pole position and is outputted to the coordinate conversion units **8** and **11** and the speed detection unit **13** so as to constitute the control system. In connection with FIG. **6** it has been explained that the phase $\theta_{ip}(2n)$ of the current difference difference vector $\Delta\Delta i(2n)$ is closer to the actual d-axis with respect to the estimated d-axis direction to which the detection use voltage V_p is applied. Namely, with FIG. **10** embodiment, even if the originally set magnetic pole position **0** is deviated from the actual d-axis, by assuming the phase $\theta_{ip}(2n)$ of the current difference difference vector $\Delta\Delta i(2n)$ as the magnetic pole position θ , the phase $\theta_{ip}(2n)$ gradually comes close to the d-axis and finally coincides with the d-axis. Once the magnetic pole position θ coincides in the controller, the controller can always and continuously detect the actual magnetic pole position, namely the d-axis.

With the present embodiment, the magnetic pole position detection can be performed through a more simplified calculation than the other calculation methods, therefore, a motor control system with more low cost can be provided. Further, such as a noise removal use low pass filter and the magnetic pole position estimation unit **16** can be added to the magnetic pole position detection unit **12** in FIG. **10**.

A further embodiment as shown in FIG. **11** through FIG. **13** relates to a method of estimating a magnetic pole position by making use of a counter electromotive force and is different from the embodiment as shown in FIG. **7** through FIG. **9**. Primary features of FIG. **11** embodiment is that a voltage gain setting unit **29** is provided and a different calculation method is used in a magnetic pole position estimation unit **30**. In the voltage gain setting unit **29**, a value of voltage gain K_p is either increased or decreased in response to the current detection pulses P_d . For example, during time from $t(2n)$ to $t(2n+1)$ K_p is set larger than 1 and during time from $t(2n+1)$ to $t(2n+2)$ K_p is set smaller than 1. Thereby, d-axis voltage command value V_{ds} and q-axis voltage command value V_{qs} respectively either increase or decrease depending on K_p . As shown in FIGS. **13A** and **13B** in vector diagrams, the voltage vector $V_a(2n)$ during time from $t(2n)$ to $t(2n+1)$ is in-phase with the voltage command vector V_s (vector sum of V_{ds} and V_{qs}) and the absolute value thereof is larger than V_s , and the voltage vector $V_b(2n+1)$ during time from $t(2n+1)$ to $t(2n+2)$ is in-phase with V_s and the absolute value thereof is smaller than V_s . When the value of K_p is, for example, set at 1.1 time and 0.9 time, an average value of $V_a(2n)$ and $V_b(2n+1)$ can be set at V_s . The current difference vectors $\Delta i_a(2n)$ and $\Delta i_b(2n+1)$ in such exemplified instance are illustrated in FIGS. **13A** and **13B**. These vectors are affected not only by the applied voltage vectors $V_a(2n)$ and $V_b(2n+1)$ but also by the counter electromotive force ($-j\omega\phi$). Thus, the phase of the counter electromotive force is detected according to the method as shown in FIG. **12** flowchart. The magnetic pole position detection unit **12** as shown in FIG. **11** is constituted by the current difference calculation unit **14** and the magnetic pole position estimation unit **30**. As has been explained in connection with FIG. **5** embodiment, the current difference calculation unit **14** calculates the current difference vectors $\Delta i_a(2n)$ and $\Delta i_b(2n+1)$, and these values are inputted into the magnetic pole position estimation unit **30** where the calculation according to FIG. **12** flowchart is performed. At first, by making use of $\Delta i_a(2n)$ and $\Delta i_b(2n+1)$ inputted at step **110** the current difference difference vector $\Delta\Delta i(2n)$ representing the difference between $\Delta i_a(2n)$ and $\Delta i_b(2n+1)$ is calculated at step **121**. Since $\Delta\Delta i(2n)$ is a difference of current difference, an influence of the counter electromotive force

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can be eliminated, and which represents current difference vector determined by $\{V_b(2n+1)-V_a(2n)\}$ of which vector relationship is shown in FIG. **13C**. At the subsequent step **122**, $\Delta\Delta i(2n)V_b/(V_b-V_a)$ is calculated to determine the current difference vector $\Delta i_{bv}(2n+1)$. The current difference vector $\Delta i_{bv}(2n+1)$ is a current difference vector when $V_b(2n+1)$ is applied under no counter electromotive force. At step **123**, a difference between $\Delta i_b(2n+1)$ and $\Delta i_{bv}(2n+1)$ and the counter electromotive force current difference vector Δi_ϕ are calculated. $\Delta i_b(2n+1)$ is a current difference vector affected by $V_b(2n+1)$ and the counter electromotive force, and $\Delta i_{bv}(2n+1)$ is a current difference vector affected only by $V_b(2n+1)$, therefore, the difference Δi_ϕ therebetween is a component affected by the counter electromotive force of which vector diagram is shown in FIG. **13D**. Accordingly, through calculation of the phase θ_ϕ of the counter electromotive force current difference vector Δi_ϕ at step **124** q-axis (negative direction thereof) can be determined. At step **113** in FIG. **12** through adding $\pi/2$ [rad] to the phase θ_ϕ d-axis namely the magnetic pole position θ is obtained.

According to the present embodiment, only through the use of the voltage gain in place of the detection use applied voltage, the magnetic pole position detection can be realized as well as through a correct drawing of vector diagram of the current different vector the phase of the counter electromotive force can be easily detected.

FIG. **14** is a control block diagram showing a further embodiment in which the characteristic of the current control system is improved by correctly detecting the counter electromotive force. The present embodiment is directed to a motor control system including a magnetic pole position sensor **50** and is intended to realize a different object than that for the control systems with no magnetic pole sensors as has been explained. Therefore, the magnetic pole position θ detected by the magnetic pole position sensor **50** is outputted such as to the coordinate conversion units **8** and **11** and to the speed detection unit **13** so as to use the motor control. Primary differences other than the above performed in FIG. **14** embodiment are those in the current control unit **7** and in a counter electromotive force detection unit **51**. In the counter electromotive force detection unit **51**, d-axis and q-axis components V_{de} and V_{qe} in the counter electromotive force are calculated from the three phase currents i_u , i_v and i_w and the magnetic pole position θ . These values are inputted into the current control unit **7** and are used for counter electromotive force compensation in the current control system, thereby, a current control characteristic such as at a time of sudden speed change can be improved. Although it is well known through addition of the counter electromotive force components in the calculation for the current control system to compensate the counter electromotive force induced inside the synchronous motor **1**, however, a method of estimating the counter electromotive force by the motor speed ω is generally used conventionally, therefore, the current frequency varies during speed variation due to excess or short compensation. Further, when the load includes a mechanical vibration system, the vibration may be amplified due to excess compensation. The present embodiment resolves these problems, and even at a time of sudden speed change the motor current can be controlled precisely along a current command value.

Details of the above will be explained with reference to FIGS. **15** and **16**. FIG. **15** is a block diagram showing processing contents performed in the current control unit **7** wherein valid/invalid of the current control system is changed over in response to the current detection pulses P_d . In FIG. **15**, a d-axis current control calculation unit **31** and

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a q-axis current control calculation unit **32** respectively feed back d-axis current and q-axis current for d-axis and q-axis current command values i_{dr} and i_{qr} and perform control calculation so as to assume their current deviations zero. In d-axis and q-axis change-over units **33** and **34**, it is changed-over whether the calculation result of the d-axis and q-axis current control calculation units **31** and **32** is outputted or zero is outputted in response to the current detection pulses P_d . More specifically, during interval from time $t(2n)$ to time $t(2n+1)$ zero is outputted and during interval from time $t(2n+1)$ to time $t(2n+2)$ the current control calculation result is outputted. In other words, when seen the above from a point as a current control system only $\frac{1}{2}$ voltage in average is outputted. Therefore, in this control system by doubling the normal gain of the current control system an equal current control characteristic can be ensured. d-axis and q-axis voltage command values V_{ds} and V_{qs} are respectively obtained by adding d-axis and q-axis counter electromotive forces V_{de} and V_{qe} calculated at a counter electromotive force detection unit **51** to the outputs of the d-axis and q-axis change-over units **33** and **34**. Now, the counter electromotive force detection unit **51** as shown in FIG. 16 will be explained. The counter electromotive force detection unit **51** is constituted by an α -axis current difference detection unit **35**, a β -axis current difference detection unit **36**, a coordinate conversion unit **37**, a d-axis counter electromotive force calculation unit **38** and a q-axis counter electromotive force calculation unit **39**. The α -axis current difference detection unit **35** and the β -axis current difference detection unit **36** are inputted of three phase currents i_u , i_v and i_w and output a current difference vector $\Delta i_a(2n)$ during interval from time $t(2n)$ to time $t(2n+1)$, in that an α -axis component $\Delta i_{a\alpha}(2n)$ and a β -axis component $\Delta i_{a\beta}(2n)$ in the current difference vector $\Delta i_a(2n)$ are detected. These current difference values in the static coordinate system (α - β axes system) are converted into ones in the d-q axes rotary coordinate system to calculate d-axis component $\Delta i_{ad}(2n)$ and q-axis component $\Delta i_{aq}(2n)$ in the current difference vector $\Delta i_a(2n)$. It is important to note that $\Delta i_{ad}(2n)$ and $\Delta i_{aq}(2n)$ are definitely d-axis and q-axis components in the current difference vector $\Delta i_a(2n)$ when seen from the static coordinate system. Further, during interval from time $t(2n)$ to time $t(2n+1)$ the outputs of the d-axis and q-axis change-over units **33** and **34** are zero, therefore, only the d-axis and q-axis counter electromotive forces V_{de} and V_{qe} are respectively outputted as the d-axis and q-axis voltage command values V_{ds} and V_{qs} . For this reason, when V_{de} is larger than the actual d-axis counter electromotive force of the synchronous motor **1**, the d-axis current difference value Δi_{ad} assumes a positive value, and contrary, when V_{de} is smaller than the actual d-axis counter electromotive force, Δi_{ad} assumes a negative value. The same is true with regard to V_{qe} . Therefore, in the d-axis counter electromotive force calculation unit **38** and the q-axis counter electromotive force calculation unit **39**, a calculation of the d-axis and q-axis counter electromotive forces V_{de} and V_{qe} is performed so that Δi_{ad} and Δi_{aq} respectively assume zero. When both Δi_{ad} and Δi_{aq} assume zero through this control calculation, it is implied that the d-axis and q-axis counter electromotive forces V_{de} and V_{qe} coincide with the actual counter electromotive forces of the synchronous motor **1**. These d-axis and q-axis counter electromotive forces V_{de} and V_{qe} are outputted to the current control unit. The above is the control method according to the present embodiment.

When a counter electromotive force is estimated from a motor speed as in a conventional art, there occurs excess or shortage between the actual counter electromotive force and

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the estimated one, and which has caused to reduce performance of the current control system. On the other hand, when the voltage which fully coincides with the actual counter electromotive force is used as the compensation amount for the counter electromotive force as in the present method, a compensation for the primary counter electromotive force is fully achieved, thereby, the performance of the current control is always kept at a high level.

FIG. 17 is a motor control system block diagram representing a further embodiment which improves the performance at a moment of motor speed sudden change in comparison with that of a conventional current control without using a magnetic pole position sensor. When comparing FIG. 17 embodiment with FIG. 14 embodiment, major differences are that the magnetic pole position sensor **50** is omitted, the detection of the magnetic pole position θ and the counter electromotive force is performed at the magnetic pole position detection unit **12** and the compensation for the counter electromotive force is performed by applying to the respective phase voltages in the static coordinate system other than the current control unit **7**. The calculation method performed in the current control unit **7** in FIG. 17 is shown in block diagram in FIG. 18 which is substantially the same as with FIG. 15 embodiment and substantially the same calculation is performed therein. Only the difference of FIG. 17 embodiment from FIG. 15 embodiment is that the counter electromotive forces V_{de} and V_{qe} are not added, therefore, the detailed explanation of the structure is omitted. FIG. 19 shows a block diagram of the magnetic pole position detection unit **12** which performs an important process in the present embodiment. At first, in a current difference detection unit **40**, respective phase current difference values $\Delta i_u(2n)$, $\Delta i_v(2n)$ and $i_w(2n)$ during interval from time $t(2n)$ to time $t(2n+1)$ are calculated in response to timing of the current detection pulses P_d . These values contain same information as in $\Delta i_{ad}(2n)$ and $\Delta i_{a\beta}(2n)$ as shown in FIG. 16. Even in FIG. 17 control system the inverter **3** is simply controlled by using respective counter electromotive forces V_{ue} , V_{ve} and V_{we} as the applied voltages during interval from time $t(2n)$ to time $t(2n+1)$, therefore, U phase, V phase and W phase counter electromotive force calculation units **41**, **42** and **43** perform calculation so that the respective current difference values $\Delta i_u(2n)$, $\Delta i_v(2n)$ and $\Delta i_w(2n)$ assume zero, of which idea is substantially the same as that for FIG. 14 embodiment. Subsequently, the phase θ_q (negative direction on q-axis) of the counter electromotive forces can be calculated based on the counter electromotive forces V_{ue} , V_{ve} and V_{we} obtained at the U phase, V phase and W phase counter electromotive force calculation units **41**, **42** and **43**, which is performed by a magnetic pole position estimation unit **44** in FIG. 19. Accordingly, with FIG. 17 system a control system can be realized which always ensures a good current control performance without a magnetic pole sensor. In the present embodiment, FIG. 17 system is constituted by a method of using a counter electromotive force, however, a method can be applied which permits both to detect the magnetic pole position by making use of the salient pole property (the inverted salient pole property) and to keep the current control property by the counter electromotive force estimation.

In the above embodiments, a method of detecting the magnetic pole position of the synchronous motor by making use of only the current sensors has been explained. As has been explained, the present invention is applicable to the synchronous motor having both a cylindrical type rotor and a rotor having a salient pole property. Further, other than the

synchronous motor, the present invention is also applicable to a reluctance motor while making use of its salient pole property. Further, although in order to avoid complexing explanation of the embodiments, an explanation on a calculation of the magnetic pole position in view of influences that the motor rotor rotates during sampling is omitted, however, such calculation can, of course, be included to the present embodiments. Further, in the present embodiments the method of detecting the magnetic pole position for every one cycle of the carrier waves has been explained, however, the detection can be realized in the same manner such as with a method of detecting the magnetic pole position by using current variation of one cycle in every plurality of cycles of the carrier waves and a method of detecting the magnetic pole position based on current variation for a unit of a plurality of cycles of the carrier waves. The present embodiments can be applied to such as an electric car and a hybrid car, besides when the present embodiments are applied to a magnet motor to which a sensorless control is now effected through an inverter control of 120° current conduction method, a sensorless system of no torque ripple and low noises can be provided through an inverter control of 180° current conduction method.

According to the present invention, only with current sensors the magnetic pole position can be detected on-line while performing the usual PWM control, thereby, a synchronous motor drive system with low noises and excellent torque control property can be provided with low cost without using a magnetic pole position sensor measuring a rotating position mechanically.

Further, in a case of using a usual magnetic pole position sensor a motor control system which always ensures a high current control performance by detecting a counter electromotive force in real time can be provided.

What is claimed is:

1. A device for controlling an AC motor, comprising:

an electric power inverter which applies a voltage to the AC motor; and

a control unit which controls the applied voltage with PWM signals in synchronism with carrier waves; wherein

the motor control device further comprises a position estimation unit which estimates a position of a rotor of the AC motor based on a detection of current in the AC motor in synchronism with the carrier waves; the phases of the applied voltages which are modified are varied depending on operating states of the AC motor.

2. A device for controlling an AC motor, comprising:

an electric power inverter which applies a voltage to the AC motor; and

a control unit which controls the applied voltage with PWM signals in synchronism with carrier waves;

wherein the motor control device further comprises a position estimation unit which estimates a position of a rotor of the AC motor through shifting the PWM signals within one cycle of the carrier waves from current variation amount of the AC motor.

3. A device for controlling an AC motor, comprising:

an electric power inverter which applies a voltage to the AC motor; and

a control unit which controls the applied voltage with PWM signals in synchronism with carrier waves;

wherein the motor control device further comprises a current variation amount detection unit which detects current variation amount of the AC motor for every

predetermined time in synchronism with the carrier waves and a position estimation unit which estimates a rotor position of the AC motor from a relationship between a plurality of the current variation amounts and a plurality of the applied voltages corresponding thereto.

4. A device for controlling an AC motor, comprising:

an electric power inverter which applies a voltage to the AC motor; and

a control unit which controls the applied voltage with PWIVL signals in synchronism with carrier waves;

wherein the motor control device further comprises a voltage addition unit which adds a detection use voltage in a direction of estimated magnetic pole position (d-axis) in synchronism with the carrier waves, a current variation amount detection unit which detects current variation amount of the AC Motor for every predetermined time in synchronism with the carrier waves and a position estimation unit which estimates a rotor position of the AC motor from a plurality of the current variation amounts.

5. A device for controlling an AC motor, comprising:

an electric power inverter which applies voltage to the AC motor; and

a control unit which controls the applied voltage with PWM signals;

wherein the motor control device further comprises a position estimation unit which estimates a position of a rotor of the AC motor by applying applied voltages having different phases in vector, a plurality of times, and by detecting counter electromotive forces from current variation amounts of the AC motor for respective voltage applications.

6. A device for controlling an AC motor, comprising:

an electric power inverter which applies voltage to the AC motor; and

a control unit which controls the applied voltage with PWM signals;

wherein the motor control device further comprises a position estimation unit which estimates a position of a rotor of the AC motor by applying applied voltages having the same phase in vector and different magnitude, a plurality of times, and by detecting counter electromotive forces from the current variation amounts of the AC motor for the respective voltage applications.

7. A device for controlling an AC motor, comprising:

an electric power inverter which applies voltage to the AC motor; and

a control unit which controls the applied voltage with PWM signals;

wherein the motor control device further comprises a position estimation unit which estimates a position of a rotor of the AC motor by controlling phases of applied voltages so that a phase difference of the current variation amounts of the AC motor with respect to the applied voltages assumes either 0° or 180°.

8. A device for controlling an AC motor, comprising:

an electric power inverter which applies voltage to the AC motor; and

a control unit which controls the applied voltage with PWM signals;

wherein the motor control device further comprises a position estimation unit which estimates a position of a

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rotor of the AC motor by controlling applied voltages so that the current variation amounts of the AC motor with respect to the applied voltages assumes 0.

9. A device for controlling an AC motor, comprising:
an electric power inverter which applies voltage to the AC motor; and
a control unit which controls the applied voltage with PWM signals;
wherein the motor control device further comprises a current variation control unit which controls the voltage so that current variation amounts of the AC motor assumes a value of 0 in a predetermined section.

10. The device according to claim 9, wherein the current variation control unit either detects or compensates counter electromotive forces of the AC motor.

11. A device for controlling an AC motor, comprising:
an electric power inverter which applies a voltage to the AC motor; and
a control unit which controls the applied voltage with PWM signals;
wherein the motor control device further comprises a phase shifting means which shifts the PWM signals within one cycle of carrier waves.

12. A device for controlling an AC motor, comprising:
an electric power inverter which applies a voltage to the AC motor; and

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a control unit which controls the applied voltage with PWM signals in synchronism with carrier waves;
wherein
the motor control device further comprises a position estimation unit which estimates a position of a rotor of the AC motor based on a detection of current in the AC motor in synchronism with the carrier waves;
and
the PWM signals are shifted so that a section for detecting current assumes a 0 voltage vector state.

13. A device for controlling an AC motor, comprising:
an electric power inverter which applies a voltage to the AC motor; and
a control unit which controls the applied voltage with PWM signals in synchronism with carrier waves;
wherein
the motor control device further comprises a position estimation unit which estimates a position of a rotor of the AC motor based on a detection of current in the AC motor in synchronism with the carrier waves;
and
the rotor position is detected by controlling applied voltages so that a phase of a current variation amount difference vector coincides with a difference vector of the applied voltage.

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